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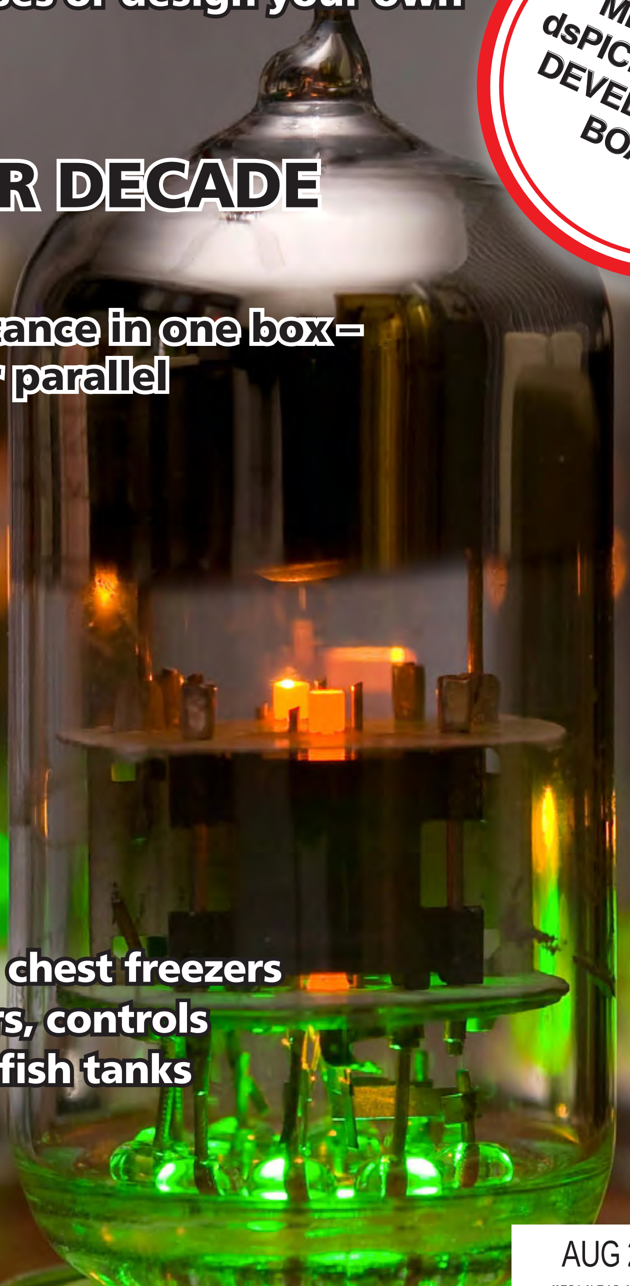

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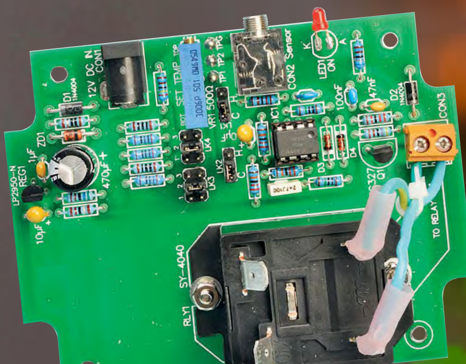

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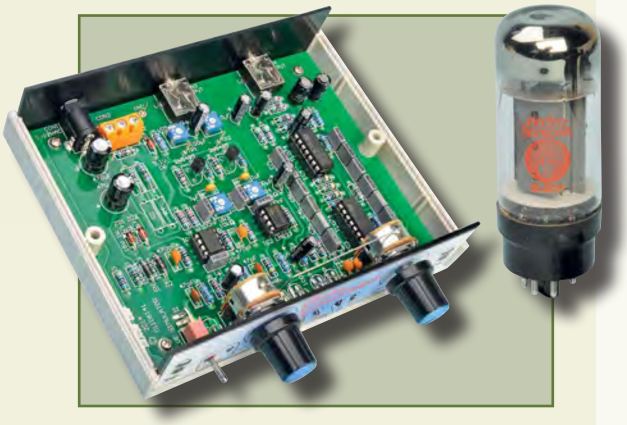

Projects and Circuits

Series and Features

Regulars and Services

Our September 2015 issue will be published on Thursday 6 August 2015, see page 72 for details.

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We have an extra-splendid issue for you this month – three superb projects and a clutch of fascinating columns to take you into the great British summer.

Valves, passive components and heat

There can be something wonderfully soothing about a valve amplifier, the distortion creates a warm sound from yesteryear, but… and it's quite a big 'but', they can be tricky and even dangerous to build for hobbyists used to the benign voltage levels that run semiconductor circuitry. I would never want to put off a careful and experienced hobbyist from tackling real valves, but why not try this month's fascinating *Nirvana Valve Sound Simulator* project? You get that special valve sound without the heat, fragility and high voltages of real valves.

Our *Resistor-Capacitor Decade Substitution Box* is one of those bits of kit every analogue experimenter should have, and once you've built it you'll never understand how you lived without it!

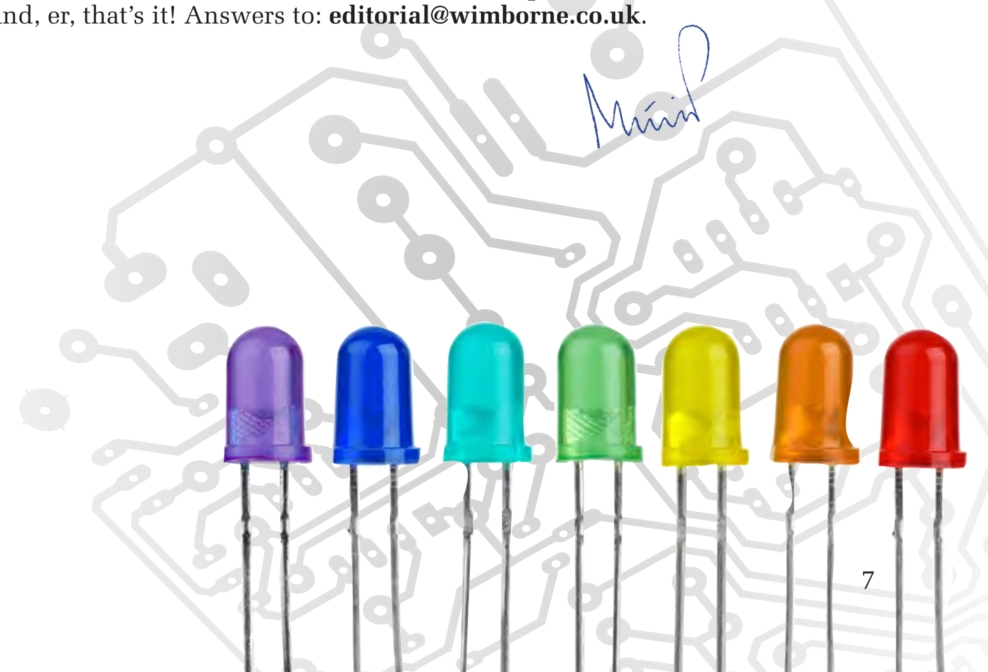
Teach-In 2015 continues to provide an excellent introduction to the nuts and bolts of building amplifiers; this month we examine heatsinks, current mirrors and more. The accompanying VU-meter project is an excellent compact design.

I don't know how they do it, but Mark Nelson and Alan Winstanley's must-read columns continue to be a fascinating distillation of the weird, wonderful and cutting-edge world of electronic technology. From liquid-metal shape-shifting antennas to Amazon's latest voice-activated gadgets, the ingenuity and elegance of modern designs never ceases to fascinate me. The pace of change is extraordinary and shows no sign of letting up.

I'm not forgetting all our other talented regulars, but they will get their day in the editorial sunshine in another issue! I do hope you enjoy this month's *EPE*, and as always, we do appreciate your email feedback.

And finally…

… we have a small music trivia competition for keen-eyed readers of *EPE*. Which one of the authors in this issue is a fan of early Pink Floyd? The first reader to let me know wins kudos, the respect of the editorial team… and, er, that's it! Answers to: **editorial@wimborne.co.uk**.



NEWS

A roundup of the latest Everyday News from the world of electronics

How to get ahead: the view from Japan – report by Barry Fox

Gone are the days when big companies with in-house R&D departments shaped trends – and launched new products only after market research told them the time was right.

'You now have to bring products to market *ahead* of market research,' Yuji Ichimura, Executive Officer, at Konica Minolta in Tokyo, said in Vienna, Austria recently. 'Market research is no longer valuable. I want to rely on the customer, not what someone else says the customer wants. We now bring products to the market while they are still developing'.

How start-ups can succeed

Ichimura was speaking during a panel discussion on 'How start-ups can profit from international corporations and vice versa', during the two-day international Pioneers Festival held annually since 2012 in Vienna's Hofburg Imperial Palace. Konica Minolta and Cisco were co-sponsoring the high-energy event, at which a mix of start-ups, hackers, and innovators from around the world pitch ideas to corporate managers and venture capitalists.

Ichimura continued, 'Start-ups want to be innovative but not get involved in legal and financial business matters. That's where large companies can help, by providing incubator services. The start-up founders can focus on their creative ideas, and the big corporations benefit from that. We are now investing in augmented reality company Wikitude. We let them focus on what they are good at, the software. That way they keep their speed.'

'When companies get bigger, different issues come up. Mr Morita (Sony), Mr Honda, Mr Matsushita (Panasonic) had passion energy and charisma. But in later generations their style didn't work out well. Strong charismatic leaders may not generate good leadership. Mr Honda is not a good manager. Not a leader. He's a crazy, technological guru. We love him! He succeeded very well. But for the following generations – the baby boomers – his style didn't work out very well.

'Once-innovative Sony has lost the market for gadgets. It now loses money on consumer electronics, but makes money on components for smartphones. Sony's movie business is so-so. PlayStation is OK, after lots of hectic periods in the past. In fact, Sony is only surviving because Mr (Kunitake) Ando started new businesses such as Sony Insurance and Sony Finance.

'Cold-calling a Japanese company, for instance by email, does not work. It has to be done by networking' Ichimura advised. 'And if a start-up comes to us and says they want to sell their company, I don't buy. If they are confident enough to grow their business, they shouldn't be selling sell their company to a large corporation. They should drive growth themselves.

Using a great idea

Ichimura cautioned, 'When you come up with a great idea, *think* there are 300 people at least in the world with the *same* idea. With energy and passion and money behind you, you can move faster than anyone else. Time is the key element of your success.'

He also warned: 'When start-ups talk to large corporations they will find some have what I call the not-developed-here (NDH) syndrome. When a large corporation is spending 5% or 10% of their revenue on R&D they have a large number of people working on new ideas. So when they talk to start-ups they will think 'Oh, OK, 10 people or 15 people in a small company can do this, we have thousands of people working on R&D, so we can do it better and faster'.

'If there is a smell of NDH in your possible partner, walk away. And make sure that you work with the business side of the company. If you are only talking to the technology managers, then your ideas have to be better than the large corporation's R&D engineers. So don't just talk to R&D – ensure you talk to the business side too.'

The road to Bletchley

The first major exhibition to explore codebreaking in World War One is now open at Bletchley Park. 'The Road to Bletchley Park' celebrates the pioneering achievements of those who waged a secret war – and how they paved the way for the codebreakers of World War Two.

The story of signals intelligence in WW1 is an untold but crucial one, because a large number of those involved went on to work with the newly formed Government Code and Cypher School (GC&CS) in 1919, which then relocated to Bletchley Park in 1939.

The first phase of this fascinating exhibition introduces the two very separate codebreaking organisations working in WW1: MI1(b), set up by the Army, and Room 40, established by the Navy. They were each fighting a secret war, behind the scenes in London offices.

For visitor information, call: 01908 640404, email: **info@bletchleypark.org.uk** or go to the Trust's website: **www.bletchleypark.org.uk**

Beta LAYOUT UV panel printing

Beta LAYOUT Ltd, a manufacturer and service provider in the prototype PCB market (PCB-POOL), has expanded its printing options for printing customised front panels.

Their new panel printer uses UV curable ink, expanding the spectrum of printable front panel materials. Front panels, which require acrylic material can now be labelled and printed to (in addition to BETA's current aluminum front panels options). For a crisp detailed print finish, white colour UV printing is also possible on acrylic material.

Formats can be printed up to A2 size with photographic quality (up to 1800 × 1800 dpi). Two of the main advantages of UV panel printing are high durability and colour brilliance.

Beta LAYOUT supports customers and the design of their custom front panels by offering free, intuitive design software – 'Front Panel Designer'. Many standardised components are contained in the software's comprehensive library and need only be selected to be included in a design. Numerous features and ordering options for mechanical processing, such as drilling holes with and without threads, flat-milling, outbreaks for fans and connectors are also available. For more details, see: **www.panel-pool.com**

Air traffic control gallery

An air traffic control gallery has opened at The National Museum of Computing (TNMOC) featuring interactive exhibits highlighting the past, present and future of air traffic control. The new gallery offers insights into the behind-the-scenes world that supports everyday air travel. It highlights the pervasiveness of computing in the modern world and how much progress has been made in a few decades.

The gallery's centre-piece is a high-fidelity air traffic control simulator

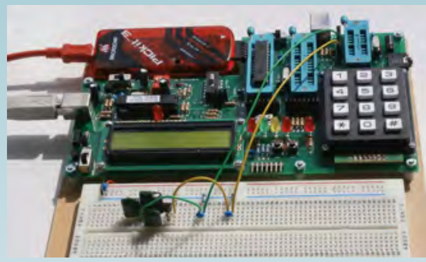
The spectacular cake that celebrated the opening of the Gallery

Visitors can experience the Museum's high-fidelity air traffic control simulator

that gives visitors a real sense of what it is like to be an air traffic controller at a control centre or major airport today. In replay mode, visitors can observe aircraft movements on a panoramic three-screen virtual airport or a control centre radar display, and listen to radio transmits between the controllers and pilots. In interactive mode, visitors can take up position at the simulators and experience, hands-on, being a controller while a member of the museum team acts as a pilot, flying the simulated aircraft in response to commands from visitors.

The gallery has an historic green-screen, round IRIS radar display (an investigative radar recording system) with working 1970s PDP-11 hardware that has been restored to working order by TNMOC volunteers.

PIC training course includes 32-bit PICs



Brunning Software have just announced their P955 PIC training course, which includes training for 32MX PICs. The first course book starts with easy-to-understand 8-bit PICs and assumes the reader has no previous programming experience. By using assembly language, the reader is given a fundamental understanding of PICs. The second book introduces PIC C, and the third book introduces serial communications between the PIC and your PC, still using 8-bit PICs. Finally, the fourth book introduces 32-bit PICs.

The key to the ease of use of this course is the P955 training circuit, which is wired to take 8-bit, 16-bit and 32-bit PICs. The P955 circuit includes a programmer for 8-bit PICs. To programme 32-bit PICs, a PICkit3 needs to be plugged onto the circuit. So, although 32-bit PICs can be difficult to understand, by starting with 8-bit PICs and using a common training circuit, an easy way to learn has been created.

During the last few chapters of PIC training the 32-bit PIC is programmed to send oscilloscope data to your PC to create a digital storage oscilloscope with advanced triggering and adjustable scan speed.

For more information, visit: **www.brunningsoftware.co.uk**

Thunderbolt 3 is go!

Intel has unveiled the latest version of its Thunderbolt 3 interface, claiming it to be the fastest, most versatile connection to any dock, display, or peripheral device – including billions of USB devices. It can deliver 40Gbps, which means one cable can handle two 4K displays.

It's not only fast, but also supplies power – up to 100W, and will use the recently launched small and reversible USB type-C connector hardware.

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TechnoTalk

Mark Nelson

People often joke about using wet string for radio antennas, but far stranger techniques have been used – and are still being devised. Prepare to have your preconceptions challenged in this foray into the unconventional with Mark Nelson.

THE IDEA THAT YOU NEED TWO conductors for transmitting signals through wire is one we were all taught, but it's not always valid. The earth makes a very good (and surprisingly low-resistance) substitute for one of the two wires and for many years telephones, and telegraphs were linked using 'half metallic' circuits consisting of one copper wire plus earth return. In other words, the earth provided one connector of the two required, with the return wire of the telephone or telegraph instruments connected to a water pipe or a so-called copper earth plate buried a few feet down in the ground. During the First World War, and afterwards, people tried connecting not one but both wires of a telephone to buried ground plates (spaced a few feet apart) at each end of the circuit and amazingly this 'wirefree' hook-up worked adequately well.

G-Line

Equally counterintuitive is G-line, which appeared in some radio theory textbooks when I first learnt radio theory to pass my amateur radio licence examination. The G stands for Goubau, one of its joint inventors, and G-line can be summed up as a single-wire transmission line that can substitute for coaxial cable and has lower loss than even the mythical Gainiax. Its main use is for conducting radio signals at UHF and microwave frequencies. Back in the 1970s it was considered highly obscure and little more than a scientific curiosity, although I do recall that it saw use at Norwood Technical College in south London, where it was employed to feed the transmitting antenna of an experimental UHF television station operated by the Royal Television Society from 1953 to around 1970. You can read a good Internet summary of how G-line works at: **http://en.wikipedia.org/wiki/Goubau_line**

Guided waves

It was as long ago as 1897 that Lord Rayleigh analysed electromagnetic-wave propagation in dielectric-filled rectangular and circular conducting tubes – or waveguides as they are now called. Patents for practical 'guided wave' radio links emerged from 1936 onwards, and by the mid-1950s the British Post Office was making serious plans for a national 'trunk waveguide' network carrying long-distance telephone calls between zone-switching centres on millimetric waves. You can think of it as microwave radio trapped inside copper pipes.

Standard Telephone Laboratories at Harlow carried considerable technical development work, although the practical and financial aspects of laying hundreds of miles of expensive copper pipe around the country delayed practical implementation (other than a 14km-long field trial in East Anglia). Eventually, commercial research into trunk waveguide communication was phased out in favour of a commitment to optical communications, a very wise move in view of the far greater practicability (and much lower cost) of optical fibres.

It's quicker by tube

But old ideas refuse to die, and the notion of guided millimetric waves has just been reinvented by a research team at the Royal University in Leuven, Belgium – with one crucial difference. Rigid and expensive copper waveguide has been replaced by flexible and far cheaper plastic tube in which the university's researchers have built a multi-gigabit communication link. Data rates up to 12.7Gbit/s and distances of up to seven metres have been achieved, using 120GHz transmitter and receiver chips with on-chip antennas and a Teflon tube that guides the signal from the transmitter to the receiver. Seven metres may not sound a fantastic range, but these are early days and for now, proof of concept is what counts.

Vast bandwidths are available and with simple modulation schemes and circuit techniques, high data rates can be achieved easily. What's more, the low complexity of the entire system results in low power consumption. In comparison with optical fibres systems, there is no electrical-magnetic interference (EMI), no excessive channel loss and no power-consuming electrical-to-optical conversion. Unlike fibres, accurate alignment of the connectors is not needed, making this solution more robust against mechanical vibrations and employing low-cost connectors. Excellent results are achieved already using widely available hollow, circular Teflon tubing with outer diameter of 2mm and (1mm inner diameter), which exhibits a loss of 2.5dB/m at 120GHz.

Solid state? Heck no!

Wet string may be a joke, but what about shape-shifting liquid radio antennas? These are for real and could be poised for commercial exploitation. That is precisely the hope of researchers at North Carolina State University in the US, who have been 'fooling' with liquid metal for over six years. They make these antennas by creating an alloy made up of the metals gallium and indium that remains in liquid form at room temperature. This is injected into very small channels the width of a human hair. The channels are hollow, like a straw, with openings at either end – but can be any shape. Once the alloy has filled the channel, the surface of the alloy oxidises, creating a 'skin' that holds the alloy in place while allowing it to retain its liquid properties.

Their latest achievement, reported by *Electronic Engineering Times*, is to construct a reconfigurable, voltage-controlled liquid metal antenna that may play a huge role in future mobile devices and the coming Internet of Things. Jacob Adams, an assistant professor at the University, explained in the paper that the researchers created the tuneable antenna so that it is controlled by voltage only by using electrochemical reactions to shorten and elongate a filament of liquid metal, thereby changing the antenna's operating frequency. Applying a small positive voltage causes the metal to flow into a capillary, while applying a small negative voltage makes the metal withdraw from the capillary.

The significance of this development is considerable, he argued. 'Mobile device sizes are continuing to shrink and the burgeoning Internet of Things will likely create an enormous demand for small wireless systems. And as the number of services that a device must be capable of supporting grows, so too will the number of frequency bands over which the antenna and RF front-end must operate. This combination will create a real antenna design challenge for mobile systems because antenna size and operating bandwidth tend to be conflicting tradeoffs.'

Build the Ni
Valve Soun

We know there are lots of valve enthusiasts out there who believe that 'valves are just better' – much more musical and pleasant to listen to than those sterile solid-state circuits with oodles of negative feedback and vanishingly small harmonic distortion. Of course, valve amplifiers do have drawbacks – they run hot and are fragile – but what if you could get 'valve sound' from a solid-state state amplifier? Well now you can, with our *Nirvana Valve Sound Simulator*.

IT WENT COMPLETELY AGAINST the grain, but we have designed a solid-state circuit which deliberately distorts. We have produce the 'desirable and musical' effects of valve amplifier circuitry in a silicon design.

OK, OK, we know that if you want genuine, true 'valve sound', the only recourse is to use a valve amplifier. But we are presenting another way to musical *nirvana*; using a solid-state amplifier with in-built valve circuitry simulation. This way, it's the valve sound you desire, but without using 'tricky' valves.

Our *Nirvana Valve Sound Simulator* can be connected in series with any solid-state mono or stereo amplifier. It can be used by musicians or in the home for normal music listening. It lets you hear what valve sound is all about, so you don't have to go to the expense of replacing a perfectly good solid-state amplifier with a valve amplifier.

What does it do?

When a valve amplifier (sometimes called a tube amplifier) is compared objectively with a modern solid-state amplifier, the results can be somewhat uncomplimentary. The valve amplifier will typically have much higher distortion, more noise, more hum and certainly a less than straight-line frequency response when driving real loudspeakers. But the sum total of those effects is what valve amplifier enthusiasts want: a mellower, softer and (it's claimed) more 'musical' sound.

Our *Nirvana Valve Sound Simulator* does not add noise and hum, but it will produce the same effects on the signal as a valve amplifier: softer symmetrical or asymmetrical clipping at the point of overload, mainly even-order harmonic distortion similar to the effects of a single pentode valve stage and a frequency response similar to that of a good quality class-AB valve amplifier with transformer coupling to the loudspeaker.

We simulate the pentode valve stage effect by using a FET source-follower in the signal path. The soft clipping effect is achieved in the same FET source-follower stage and it is fully adjustable for degree and asymmetry.

We also need to simulate the effect of a valve amplifier's output impedance on the frequency response of a typical loudspeaker. This is where solid-state amplifiers have a big advantage over valve amplifiers. Well-designed solid-state amplifiers usually have an output impedance which is less than one-sixtieth of the nominal impedance of a loudspeaker, ie, something less than 150mΩ. By contrast, a valve amplifier typically has an output impedance of about 2Ω, depending on how much negative feedback is applied from the output back to the earlier stages.

The relatively high output impedance of the valve amplifier has two effects when driving loudspeakers. The first effect is a much lower 'damping factor', which is the ratio of nominal loudspeaker impedance to the amplifier's output impedance. For a solid-state amplifier, we expect to see damping factors of 60 or more and that means that the amplifier has very tight control over the movement of the loudspeaker cone. This leads to less 'boomy' bass and lower distortion of bass frequencies.

An equally important advantage of a very low output impedance is a much more linear frequency response from all loudspeakers. This is because loudspeakers do not have a constant impedance, but one that varies widely with frequency. So ideally, a loudspeaker should be driven from a voltage source and that means having a low output impedance amplifier.

With the much higher output impedance of a valve amplifier, the considerable variations in a loudspeaker's impedance over the frequency range means that the overall response will be much 'bumpier' or less smooth. Say, for example, a valve amplifier has an output impedance of 2Ω and the loudspeaker has a nominal output impedance of 8Ω. That means that 25% of the drive signal will be lost within the amplifier itself.

But the effect is much worse because the loudspeaker's impedance varies from less than 6Ω to more than 30Ω.

Fig.1 shows a bass reflex loudspeaker impedance curve. Typically, these have a double hump at low frequencies. They may rise to a second broad peak at the mid-frequencies (depending on the effect of the crossover network) and then rise again at the high end, due to the inductance of the tweeter. By contrast, loudspeakers in sealed cabinets have only one peak at the low frequency end.

Any increase in impedance above the nominal value (eg, 8Ω) at a particular frequency will result in a boost to the loudspeaker's response, while a reduction will result in a drop in the response – see Fig.3. This diagram depicts the effect on the frequency response of four loudspeaker systems, (originally published in *Silicon Chip* magazine), when driven by an amplifier with an output impedance of 4Ω.

As you can see, the main areas of boosting occur at the two bass resonances and at the mid-band impedance hump. For example, with the JV100 loudspeaker depicted at the top of Fig.3, the boost is as much as +3.9dB. Similarly, there is a broad boost to the response of more than +3dB from around

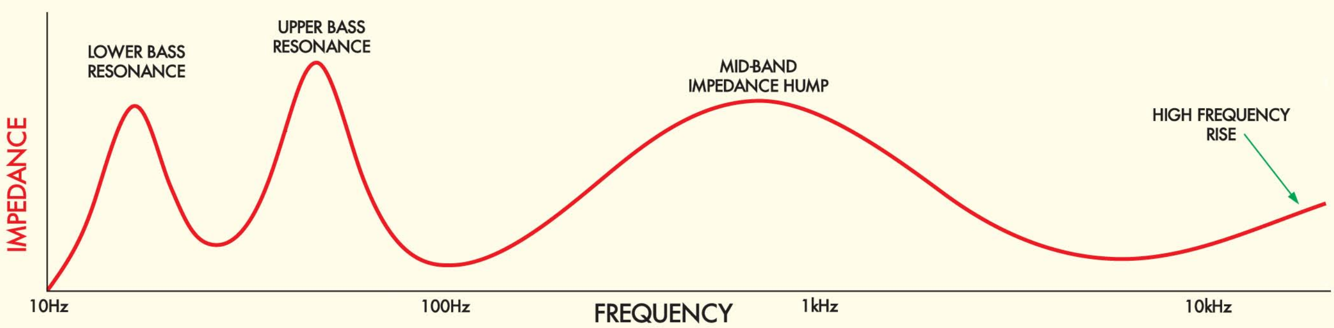
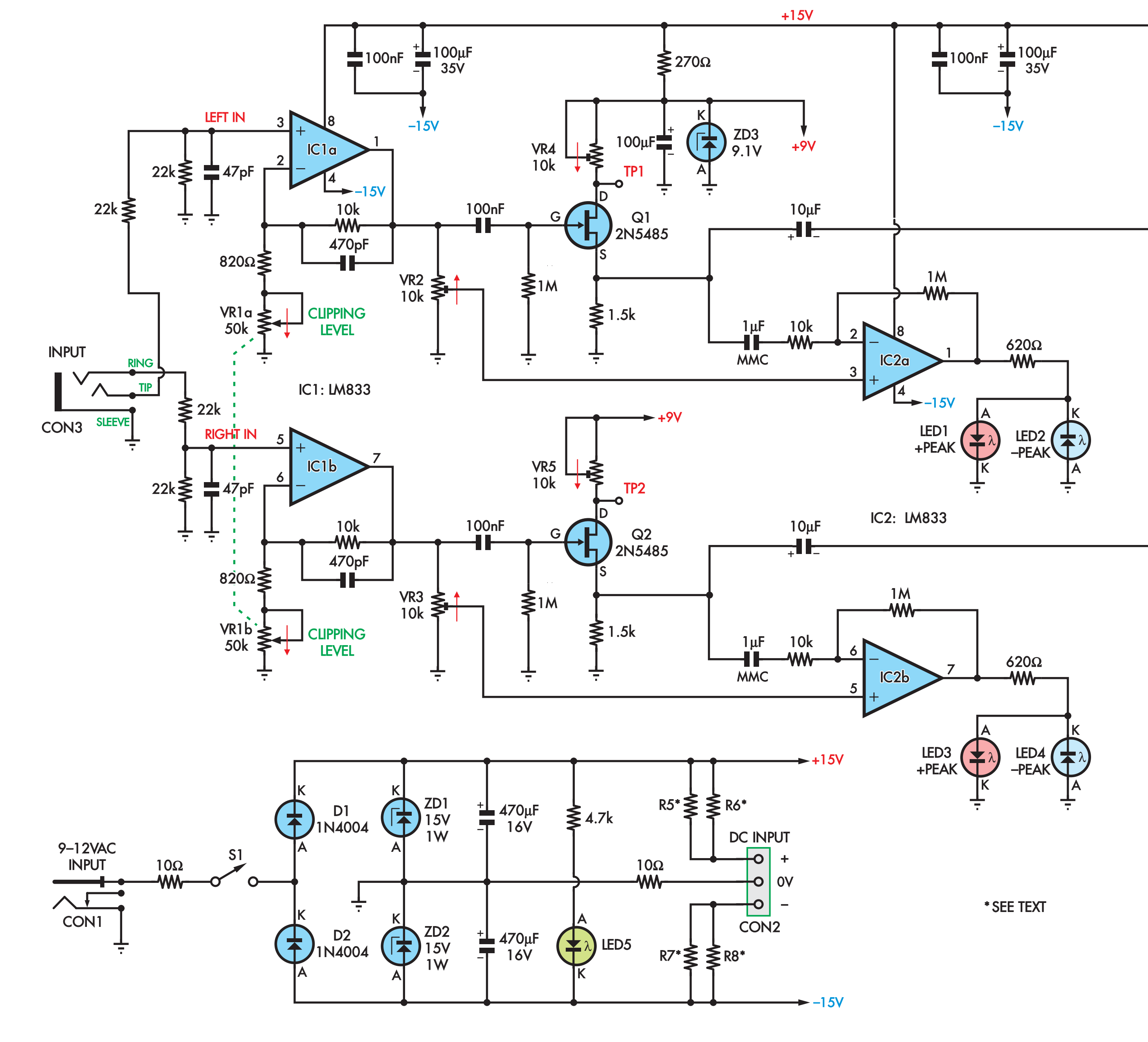


Fig.1: a typical bass reflex loudspeaker impedance curve. As shown, there's a double hump at low frequencies, with the impedance then rising to a broad peak at the mid-frequencies (depending on the effect of the crossover network) and then rising again at the high end, due to the inductance of the tweeter.



NIRVANA VALVE SOUND SIMULATION

500Hz to 1.5kHz and a smaller boost to the tweeter at the high-frequency end.

By contrast, if the same loudspeaker is driven by a solid-state amplifier with a typical output impedance of less than 150mΩ, there is no boost or cut, as it should be!

The *Nirvana* simulates these loudspeaker frequency deviations with a number of individually adjustable filters that are varied by the 'Loudspeaker Response' control. The selection of a particular loudspeaker for simulation requires choosing a particular set of component values – to be discussed later in this article.

The other control on the front panel of the *Nirvana Valve Sound Simulator* is for 'Clipping Level'.

If you want to delve more into valve sound, here are some interesting sites:

1) **http://spectrum.ieee.org/consumer-electronics/audiovideo/the-cool-sound-of-tubes**
2) **http://spectrum.ieee.org/consumer-electronics/audiovideo/the-cool-sound-of-tubes/distortion**
3) **http://en.wikipedia.org/wiki/Tube_sound**

In use, the *Nirvana Valve Sound Simulator* connects between the preamplifier outputs and the power amplifier inputs of a solid-state amplifier. In amplifiers with a tape loop you can use this facility, while for a musician's (eg, guitar) amplifier, it would be connected into the effects loop.

As shown in the photos, the unit is housed in a compact case and can be powered from an AC plugpack. Alternatively, balanced DC supply rails

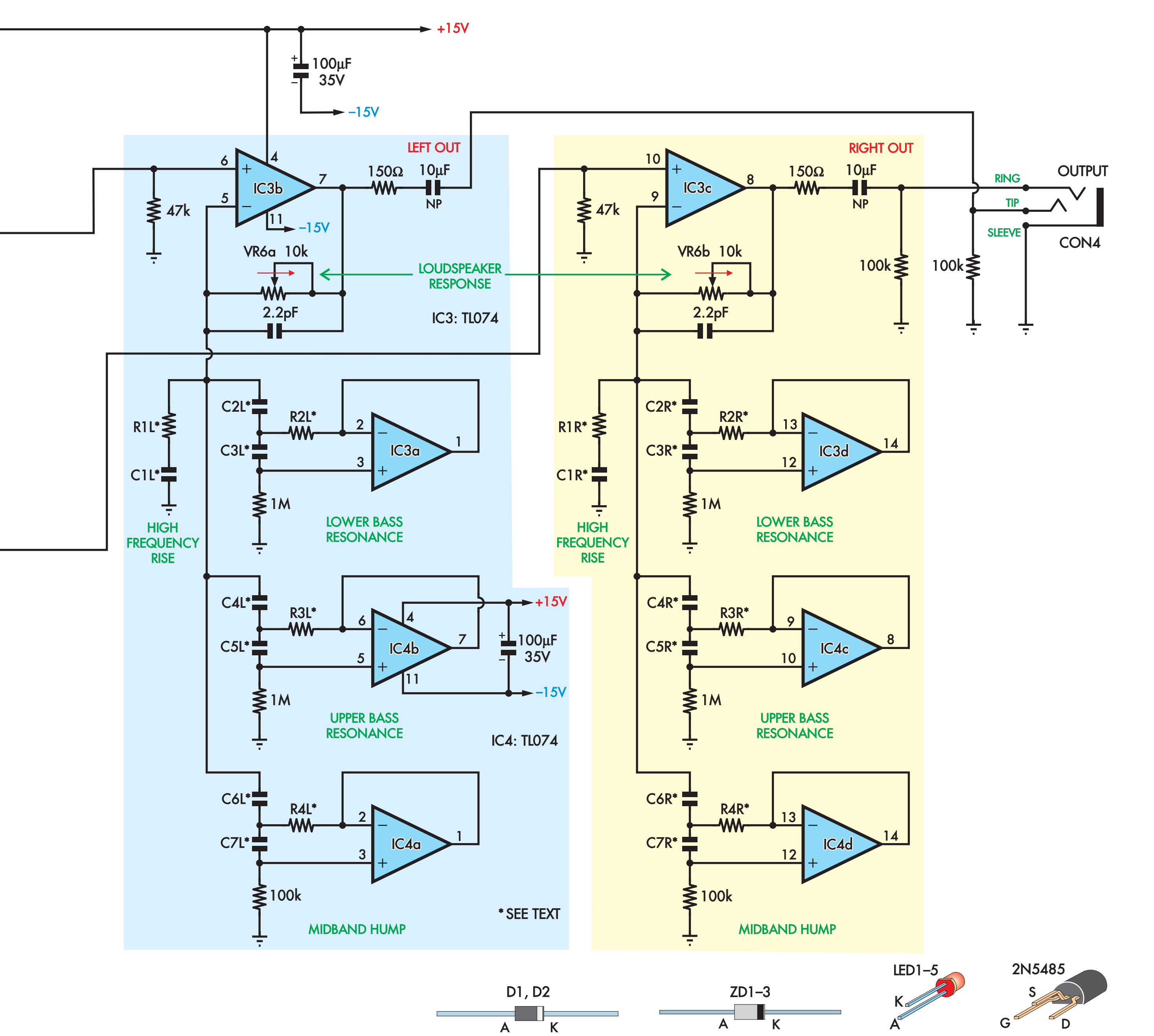


Fig.2: the complete circuit of the *Nirvana Valve Sound Simulator*. The input signals from CON3 are amplified by IC1, then distorted and clipped by JFETs Q1 and Q2. IC2 provides an indication of clipping symmetry while IC3 and IC4 act as parametric equalisers to adjust the frequency response to match that of a typical valve amplifier driving loudspeakers.

could be obtained from existing equipment. The socket for the AC supply is accessed from the rear, as are the 3.5mm stereo input and output sockets.

Circuit details

Refer now to Fig.2 for the circuit details. Each channel uses six op amps (all in four ICs) and a JFET, and both channels are identical.

The input signal is applied via CON3, a stereo 3.5mm jack socket. If only a mono signal is required, then a mono jack plug can be used to apply signal to the left channel only. This will connect the ring terminal to ground and so prevent signal in the right channel.

The following circuit description is for the left channel signal path. As shown, signal is applied via the tip connection of CON3 and is reduced by a factor of two, using two 22kΩ resistors, so that line-level signals will not necessarily cause clipping in the following JFET stage if op amp IC1a is set for minimum gain.

IC1a's gain can be varied between 1.2 and 13 by potentiometer VR1a, which sets the signal clipping level in the JFET stage. When VR1a is set for minimum gain, the input signal needs to reach 1.66V RMS before clipping occurs and when VR1a is set for maximum gain, the input signal only needs to reach 109mV RMS before clipping.

Following IC1a is the JFET amplifier stage, Q1. This is configured as a source follower (similar to a bipolar transistor emitter-follower or a valve cathode-follower). The JFET produces harmonic distortion similar to that in pentode valve stages (predominantly

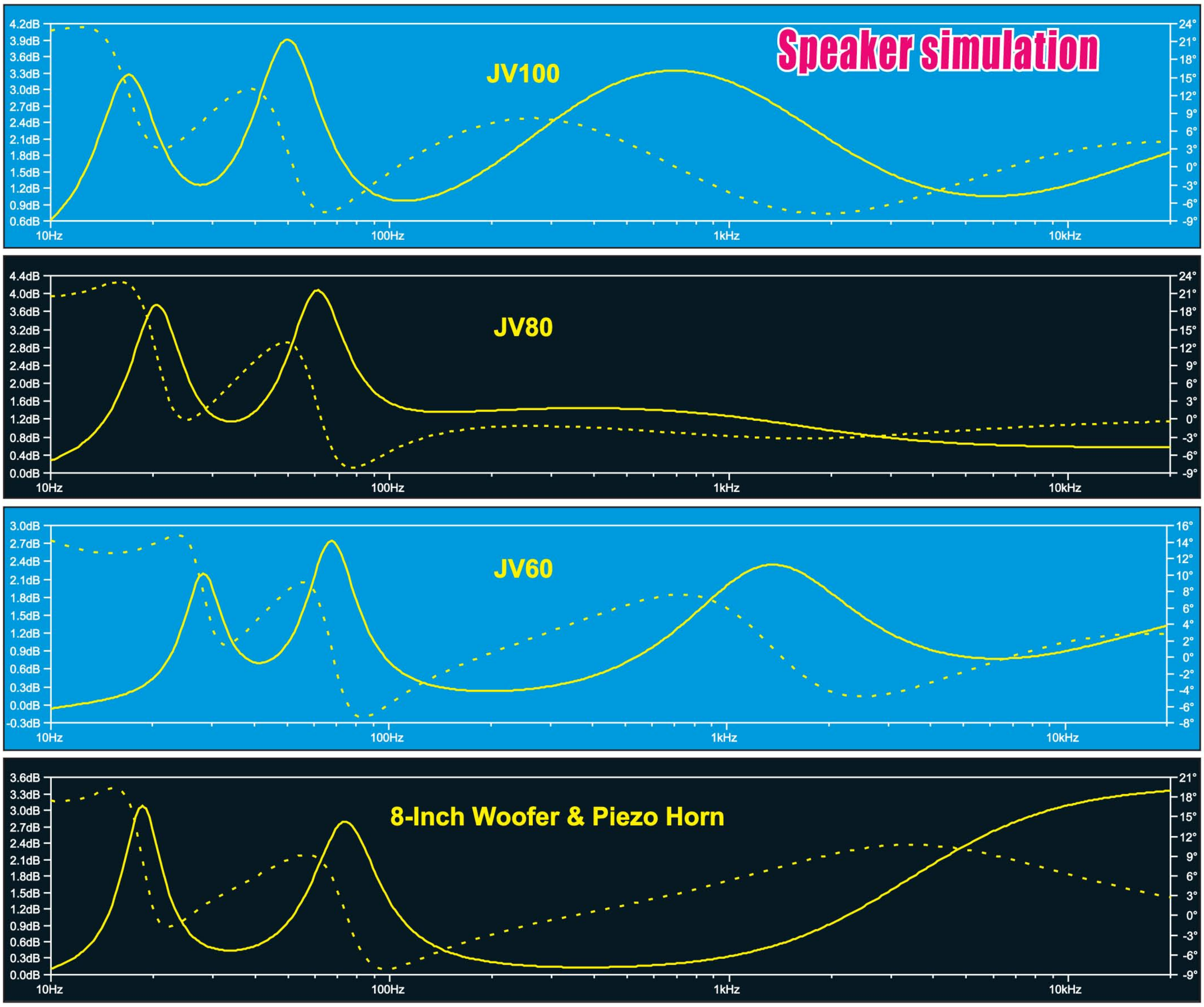


Fig.3: these curves simulate the wide deviations from a flat frequency response for four loudspeakers previously published in *Silicon Chip*, caused by the interaction of the varying loudspeaker impedance with the typical 4Ω output impedance of a valve amplifier. The amount of boost can be seen on the left-hand vertical axis (in dB) while the deviation in phase is shown in the dotted curves and the corresponding right-hand vertical axis (in degrees). These same effects can be simulated with the Loudspeaker Response control of the *Nirvana Valve Sound Simulator*.

even harmonics) and it also produces soft signal clipping when overloaded.

The signal from IC1a is fed to Q1's gate via a 100nF capacitor, while the signal output is taken from Q1's source. Trimpot VR4 adjusts Q1's operating current, which varies the symmetry of clipping – whether the signal clips symmetrically – or, it clips the positive or negative signal swings more severely.

IC2a drives the positive and negative clipping indicators. It compares the input and output signals of Q1. When the signals differ, such as when Q1 is clipping, the output of IC2a swings high or low to drive LED1 (positive clipping) or LED2 (negative clipping). For this indication to be accurate, IC2a's gain needs to be carefully adjusted to be equal to the gain of Q1, using trimpot VR2 (or VR3 in the right channel).

Main Features

- Simulates the frequency response of a valve amplifier when driving loudspeakers
- Provides mainly even-ordered harmonic distortion, ie, second, fourth, sixth...
- Input level control sets distortion threshold and clipping
- Soft clipping on overload
- Clipping indicators for positive and negative signal excursions
- Clipping symmetry can be adjusted
- One of four different loudspeaker responses can be used or design your own
- Can run from a 9-12VAC supply (eg, a plugpack) or a ±12VDC to ±45V DC dual supply (eg, from existing equipment)

Loudspeaker simulation

The output signal from Q1 is then fed to the loudspeaker simulator section, which comprises op amps IC3b, IC3a,

IC4b and IC4a (the equivalent functions in the right channel are provided by IC3c, IC3d, IC4c and IC4d).

IC3b can be regarded as the main op amp, and its feedback network is modified by op amps IC3a, IC4a and IC4b, which can each be regarded as single-frequency equalisers, much like those used in gyrator-based graphic equalisers. The difference is that we have no slider controls to vary the individual equalisers. The maximum gain at high frequencies is set by 'high-frequency rise' components R1L and C1L, and the overall gain is set by VR6a, the Loudspeaker Response control.

IC3a is the equaliser providing the simulated lower frequency impedance peak in a bass-reflex loudspeaker system. IC4b adds the upper bass peak for bass-reflex systems and the main peak in sealed systems. In the latter case, IC3a is effectively disabled and has no effect on the overall frequency response.

Finally, IC4a provides a mid-band impedance hump that may be present with some speaker systems.

So each of the three equalisers boosts a defined frequency band about a certain centre frequency.

By selecting the values of the capacitors and resistors, we can set the required tuning frequency and shape of the boost. We designed the speaker impedance simulation circuitry using LTSpice (see **www.linear.com/designtools/software**). This SPICE simulation program from Linear Technology can be used with Windows or Mac operating systems.

The circuit file for this loudspeaker simulation (**Valve Simulator.asc**) is available on the *EPE* website. You can change the values and set the loudspeaker simulation curve yourself if you wish. Otherwise, we have a table that produces impedance curves for some typical loudspeakers.

Parts List

1 double-sided PCB, available from *EPE PCB Service*, code 01106141, 129.5 × 100mm
1 front-panel PCB, available from *EPE PCB Service*, code 01106142
1 ABS instrument case, 140 × 110 × 35mm
1 9-12V 50mA AC plugpack (optional, see text)
1 PCB-mount DC socket (CON1)
1 3-way PCB-mount screw terminal block, 5.08mm pitch (CON2)
2 3.5mm PCB-mount stereo jack sockets (CON3,CON4)
1 SPDT PCB-mount toggle switch (S1) (Altronics S 1421)
1 16mm dual-gang 50kΩ linear potentiometer (VR1)
1 16mm dual-gang 10kΩ linear potentiometer (VR6)
4 10kΩ horizontal trimpots (VR2-VR5)
2 knobs to suit potentiometers
2 DIL8 IC sockets (optional)
2 DIL14 IC sockets (optional)
4 No.4 × 6mm self-tapping screws
4 PC stakes (GND,GND,TP1,TP2)
1 100mm length of 0.7mm tinned copper wire

Semiconductors

2 LM833 op amps (IC1,IC2)
2 TL074 quad op amps (IC3,IC4)
2 2N5485 JFETs (Q1,Q2)
2 3mm high-intensity red LEDs (LED1,LED3)
2 3mm high-intensity blue LEDs (LED2,LED4)
1 3mm high-intensity green LED (LED5)
2 15V 1W zener diodes (ZD1,ZD2)
1 9.1V 1W zener diode (ZD3)
2 1N4004 1A diodes (D1,D2)

Capacitors

2 470µF 16V PC electrolytic
5 100µF 35-63V PC electrolytic
2 10µF 16V PC electrolytic
2 10µF 16V NP PC electrolytic
2 1µF monolithic ceramic
4 100nF MKT
2 470pF ceramic
2 47pF ceramic
2 2.2pF ceramic

Selected capacitors

JV100 simulation: 2 × 330nF, 2 × 150nF, 2 × 47nF, 2 × 22nF, 2 × 6.8nF, 2 × 1nF MKT, plus 2 × 470pF ceramic
JV80 simulation: 2 x 270nF, 2 × 100nF, 2 × 56nF, 2 × 22nF, 2 × 6.8nF, 2 × 1nF MKT
JV60 simulation: 2 x 120nF, 2 × 82nF, 2 × 22nF, 2 × 12nF, 2 × 6.8nF, 2 × 1nF MKT, plus 2 × 470pF ceramic
8-inch woofer with piezo horn simulation: 2 x 270nF, 2 × 100nF, 2 × 33nF, 4 × 4.7nF MKT

Resistors (0.25W, 1%)

8 1MΩ	2 1.5kΩ
4 100kΩ	2 820Ω
2 47kΩ	2 620Ω
4 22kΩ	1 270Ω
4 10kΩ	2 150Ω
1 4.7kΩ	2 10Ω

Selected resistors

JV100 simulation: 2 × 22kΩ, 4 × 12kΩ, 2 × 10kΩ
JV80 simulation: 2 × 33kΩ, 4 × 10kΩ
JV60 simulation: 2 × 22kΩ, 4 × 12kΩ, 2 × 10kΩ
8-inch woofer with piezo horn simulation: 2 × 10kΩ, 4 × 8.2kΩ

Power supply resistors

R5-R8: see text and Table 1

Power supply

Power for the circuit can come from an AC plugpack (9-12V) rated at 50mA or more. Alternatively, positive and negative DC supply rails from existing equipment can be used. In the latter case, power is applied via CON2.

Resistors R5, R6, R7 and R8 are used when the external supply is 15V or more. They provide the voltage drop for 15V zener diodes ZD1 and ZD2. Table 1 on the following page shows the resistor values required for various supply voltages.

Construction

Construction is straightforward, with all the parts mounted on a PCB, available from the *EPE PCB Service*, coded 01106141 and measuring 129.5 × 100mm. This is housed in a small instrument case measuring 140 × 110 × 35mm (W × D × H).

Before installing any of the parts, you need to use Table 2 to select the required values for resistors R1-R4 and capacitors C1-C7 to simulate a particular speaker. These values depend on the speaker load that is being simulated, as explained earlier.

Basically, Table 2 shows the values required to simulate various loudspeaker loads. In other words, you can simulate the sound of a valve amplifier driving one of these types of speakers.

If you don't have a preference, we suggest using the JV80 values. Alternatively, you can determine your own

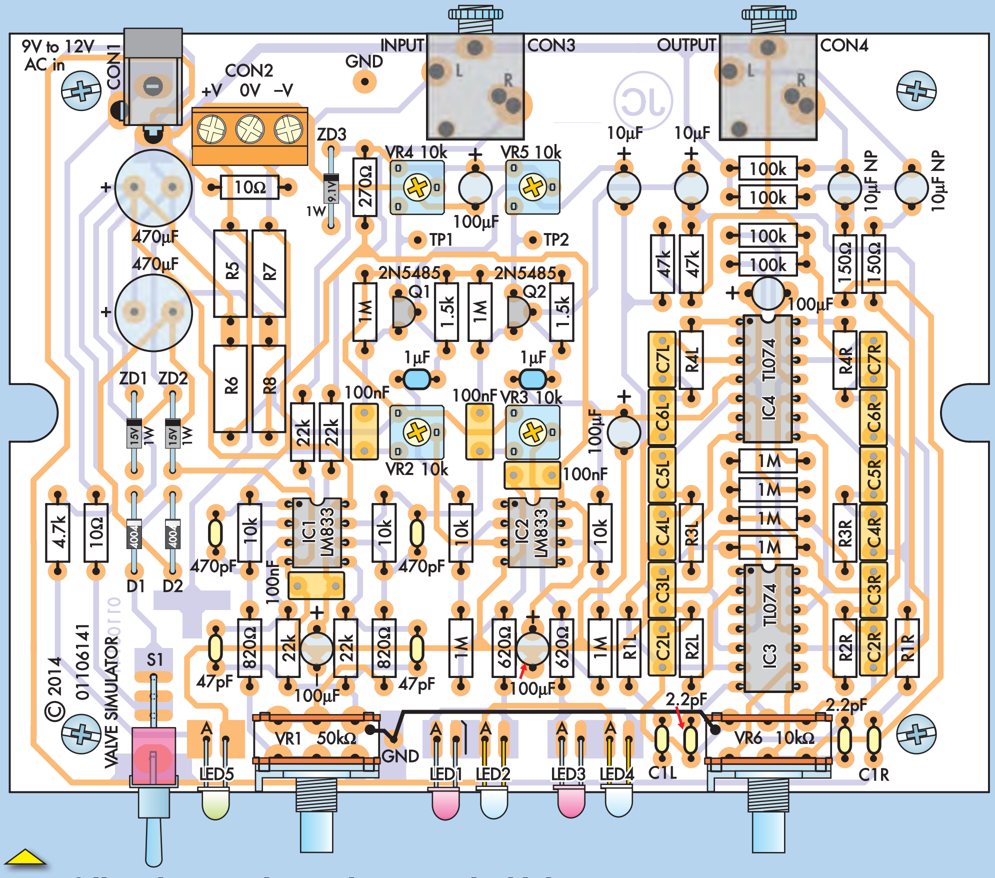


Fig.4: follow this parts layout diagram to build the PCB. Resistors R1-R4 and capacitors C1-C7 in the filter networks are selected from Table 2, while the power supply resistors (R5-R8) are selected from Table 1 (see text).

The PCB is fastened into the case using four self-tapping screws which go into integral corner pillars.

Table 1. Dropping resistors for external dual supply rails

Supply voltage	R5	R6	R7	R8
±45VDC	2.7kΩ 1W	2.7kΩ 1W	2.7kΩ 1W	2.7kΩ 1W
±40VDC	2.2kΩ 1W	2.2kΩ 1W	2.2kΩ 1W	2.2kΩ 1W
±35VDC	1.5kΩ 1W	1.5kΩ 1W	1.5kΩ 1W	1.5kΩ 1W
±30VDC	620Ω 1W	–	620Ω 1W	–
±25VDC	390Ω 1W	–	390Ω 1W	–
±20VDC	220Ω 1/2W	–	220Ω 1/2W	–
±15VDC	10Ω 1/2W	–	10Ω 1/2W	–
±12VDC	10Ω 1/2W	–	10Ω 1/2W	–

Note: a dash (–) means that no component is installed.

component values based on LTSpice simulation, as explained earlier.

You also need to decide on the power supply that you will be using and select resistors R5-R8 from Table 1 if using an external split DC supply (ie, one with positive and negative supply rails). This could come from a power amplifier or preamplifier, for example.

Alternatively, resistors R5-R8 are not required if using an external 9-12VAC plugpack supply.

Fig.4 shows the parts layout on the PCB. Begin the assembly by installing the resistors. Table 3 shows the resistor colour codes, but you should also

Table 2: R and C values for vented, sealed and piezo horn loudspeakers

Loudspeaker	VR6 Setting	HF rise		First impedance peak			Second impedance peak			Midband hump		
		C1	R1	C2*	C3*	R2*	C4	C5	R3	C6	C7	R4
JV100 (8Ω)	5.6kΩ	470pF	22kΩ	330nF	22nF	12kΩ	150nF	6.8nF	10kΩ	47nF	1nF	12kΩ
JV80 (8Ω)	5.6kΩ	–	–	270nF	22nF	10kΩ	100nF	6.8nF	10kΩ	56nF	1nF	33kΩ
JV60 (4Ω)	3.9kΩ	470pF	22kΩ	120nF	22nF	12kΩ	82nF	6.8nF	10kΩ	12nF	1nF	12kΩ
8-inch speakers, with piezo horn (8Ω)	3.9kΩ	4.7nF	8.2kΩ	270nF	33nF	8.2kΩ	100nF	4.7nF	10kΩ	–	–	–

Note 1: R and C numbers show an 'L' suffix for the left channel components and an 'R' suffix for the right channel components on the circuit and PCB layout. Note 2: * denotes no component for a sealed enclosure. Note 3: VR6 setting shown is for 4Ω output impedance amplifiers. VR6 is set to a lower resistance for lower output impedance. Note 4: a dash (–) means that no component is installed.

check each one using a DMM before mounting it in place.

Follow with the IC sockets, diodes D1 and D2, zener diodes ZD1-ZD3 and trimpots VR2-VR5. Take care to ensure that the diodes and zener diodes are oriented correctly and note that the IC sockets all face in the same direction (ie, pin 1 at top left).

The capacitors are next on the list. Table 4 shows the codes used on the smaller ceramic and MKT types. Be sure to orient the polarised electrolytic types correctly and note that the two 10µF electrolytics at top right are non-polarised (NP).

Switch S1 and power socket CON1 are necessary only if using the AC plugpack for the supply. Conversely, 3-way screw terminal block CON2 is necessary only if you are using an external split DC supply.

Now for the two potentiometers (VR1 and VR6). Before fitting them, cut their shafts to suit the knobs using a hacksaw and clean up the ends with a file. It's also necessary to file away a small area of the passivation layer at the top of each pot body, to allow an earth wire to be soldered in place later (see Fig.4).

The pots are then fitted to the PCB, noting that VR1 is 50kΩ and VR6 is 10kΩ. Push them all the way down onto the PCB before soldering their pins.

The two 3.5mm jack sockets (CON3 and CON4) can go in next, followed by PC stakes for TP1 and TP2 and at the two GND positions (one to the right of VR1 and one to the left of CON3).

Installing the LEDs

The five LEDs are installed with their leads bent down through 90°, so that they later protrude through matching holes in the front panel. First, check that the anode (longer) lead is to the left (lens facing towards you), then bend both leads down through 90° exactly 8mm from the rear of the plastic lens. This is best done by folding them over a cardboard strip cut to 8mm wide.

Once that's done, install each LED so that its horizontal leads are exactly 4mm above the PCB. In practice, it's just a matter of pushing each LED down onto a 4mm-thick spacer (eg, a cardboard strip) before soldering its leads. Use a green LED for LED5, red LEDs for LEDs1 and 3 and blue LEDs for LEDs 2 and 4.

Table 4: Capacitor Codes

Value	µF Value	IEC Code	EIA Code
1µF	1µF	1u0	105
100nF	0.1µF	100n	104
470pF	NA	470p	471
47pF	NA	47p	47
2.2pF	NA	2p2	2.2

Table 3: Resistor Colour Codes

❏	No.	Value	4-Band Code (1%)	5-Band Code (1%)
❏	8	1MΩ	brown black green brown	brown black black yellow brown
❏	4	100kΩ	brown black yellow brown	brown black black orange brown
❏	2	47kΩ	yellow violet orange brown	yellow violet black red brown
❏	4	22kΩ	red red orange brown	red red black red brown
❏	4	10kΩ	brown black orange brown	brown black black red brown
❏	1	4.7kΩ	yellow violet red brown	yellow violet black brown brown
❏	2	1.5kΩ	brown green red brown	brown green black brown brown
❏	2	820Ω	grey red brown brown	grey red black black brown
❏	2	620Ω	blue red brown brown	blue red black black brown
❏	1	270Ω	red violet brown brown	red violet black black brown
❏	2	150Ω	brown green brown brown	brown green black black brown
❏	2	10Ω	brown black black brown	brown black black gold brown

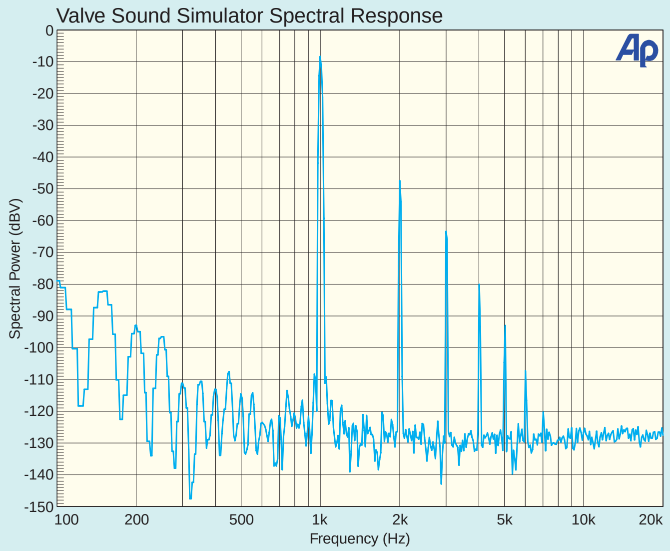


Fig.5: spectrum analysis of the output signal (1kHz input), showing strong second harmonic distortion along with third, fourth, fifth and sixth harmonics at lower levels.

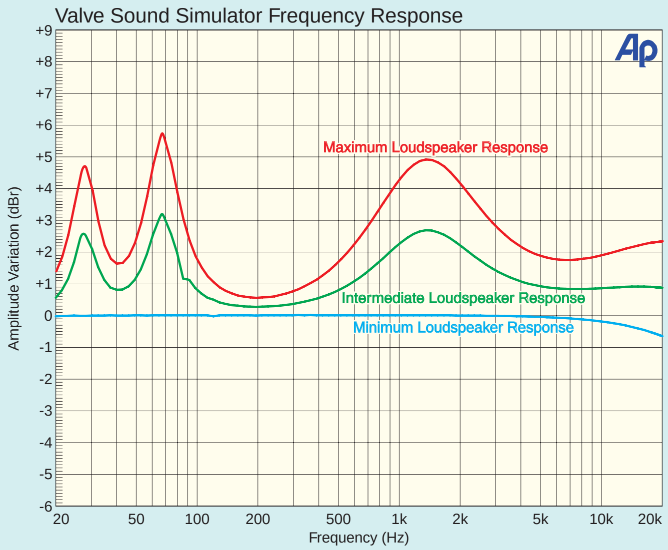


Fig.6: this graph shows the frequency response of the unit when set to simulate driving JV60s, with the Loudspeaker Response knob in three different positions.

The PCB assembly can now be completed by earthing the pot bodies to the GND PC stake next to VR1. That's done using a length of 0.7mm-diameter tinned copper wire (see Fig.4 and photos). You can straighten the tinned copper wire by clamping one end in a vice and then stretching it slightly by pulling on the other end with pliers. It can then be bent to shape so that it contacts the GND stake and is soldered.

Final assembly

Before installing the PCB assembly in the case, you have to drill a number of holes for the front and rear panels. The accompanying panel artworks (Fig.7) can be copied and used as drilling templates.

On the front panel, you will need to drill (and ream) a 5mm hole for switch S1, 3mm holes for LEDs1-5 and 7mm holes for the pot shafts. The two stereo jack sockets on the rear panel require 6mm holes, while the DC power socket requires a 6.5mm access hole.

Once that's done, print the artworks from the website onto photo paper and attach them to the panels using silicone sealant. The holes can then be cut out with a sharp hobby knife.

Alternatively, you can purchase a PCB-based front panel (blue with white labels) with pre-drilled holes from the *EPE PCB Service*.

After that, it's just a matter of fitting the panels to the PCB, sliding the assembly into the case and securing the PCB to the four corner mounting pillars using No.4 self-tapping screws. The assembly can then be completed by pushing the knobs onto the pot shafts. Reposition the end pointers of the knobs if necessary, so that they correctly point to the fully anti-clockwise and fully clockwise positions.

Testing

If you haven't already done so, insert the four ICs into their sockets, taking care to orient them correctly. Next, apply power and check that the power LED lights. If that checks out, check the supply voltage between pins 8 and 4 of both IC1 and IC2 and between pins 4 and 11 of IC3 and IC4. This should be around 30V DC if you are applying 12VAC via CON1. Alternatively, you can apply ±12V DC or more via 3-way screw terminal block CON2.

Note that you will only get around 25V (ie, ±12.5V) if using a 9VAC supply. Regardless, there should be about 9.1V across ZD3.

Assuming these supply voltages are all correct, follow this step-by-step procedure to adjust the unit:

Step 1 Connect a DMM set to volts between TP1 and a GND stake and adjust VR4 for a reading of

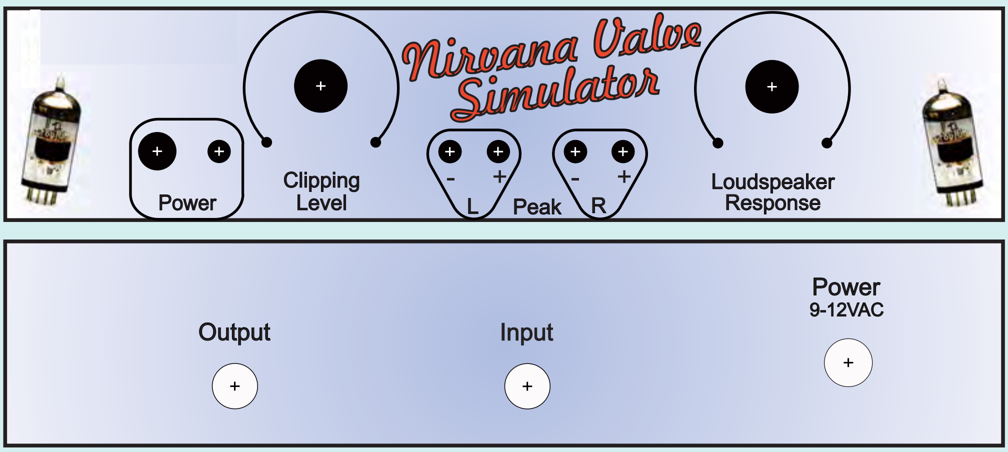


Fig.7: these two artworks can be copied and used as drilling templates for the front and rear panels.

5.8V. Similarly, adjust VR5 for a reading of 5.8V at TP2. This gives more or less symmetrical clipping for both Q1 and Q2.

Step 2 Apply a low-level 1kHz signal to both the left and right inputs and adjust VR2 and VR3 so that the positive and negative peak LEDs in both channels are off. You will find that there's a 'dead spot' in each trimpot's setting range where both LEDs are off. Set each trimpot to the middle of its dead spot.

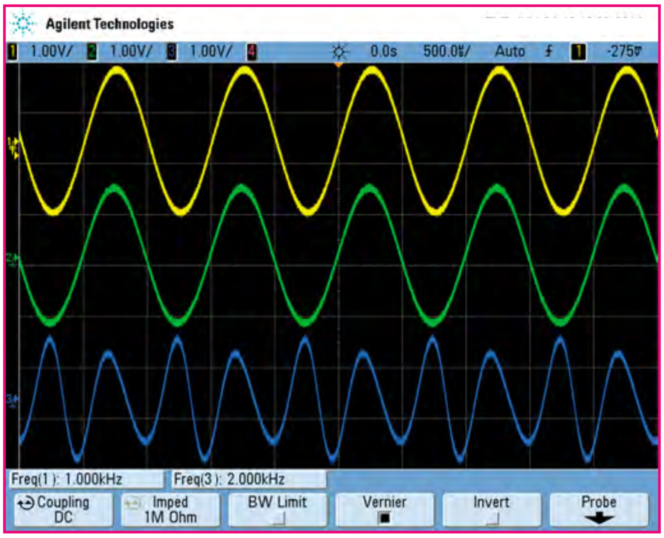
If the LEDs do not extinguish with this adjustment, try reducing the signal level using VR1 or at the signal generator (note: if you don't have a signal generator, it's easy to find a virtual instrument online).

Step 3 Increase the signal level so that the clipping LEDs begin to light. When that happens, readjust trimpots VR4 and VR5 to give symmetrical clipping, so that both the red and blue clipping LEDs light at the same time (ie, for the positive and negative signal excursions).

Finally, note that the input and output sockets can be linked to RCA connectors via adaptor cables (ie, 3.5mm stereo jack plug to RCA). For mono use, a mono 3.5mm jack plug can be used, in which case only the left channel will be supplied with signal and the right channel input will be grounded. A mono plug could then also be used for the output since the right channel will not have any output.

The rear panel carries access holes for the input and output sockets and for the power socket. Note how the metal bodies of the two pots are earthed to the GND stake using a length of tinned copper wire.

Fig.8: the output of the unit (green) compared to the input (yellow) at 1kHz. The signal level is set below clipping and the distortion residual (blue) is primarily second harmonic. This can be clearly seen as the residual is at twice the fundamental frequency, ie, 2kHz.



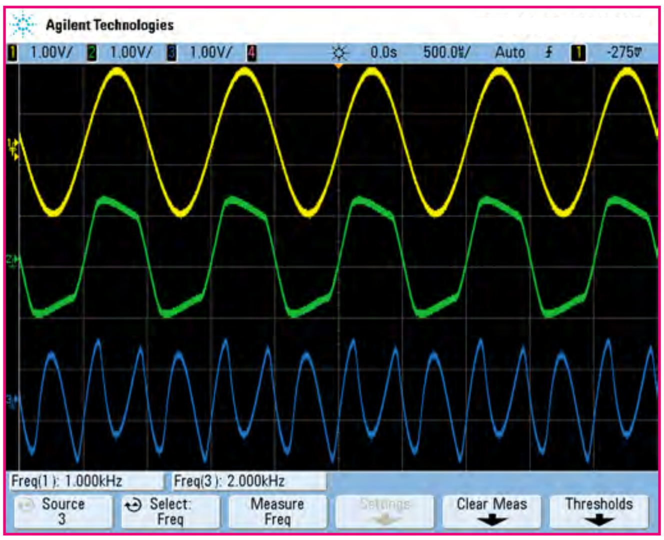


Fig.9: the same traces as in Fig.8 but with more input signal, causing clipping. The effects of soft clipping and the frequency response shaping filter are evident.

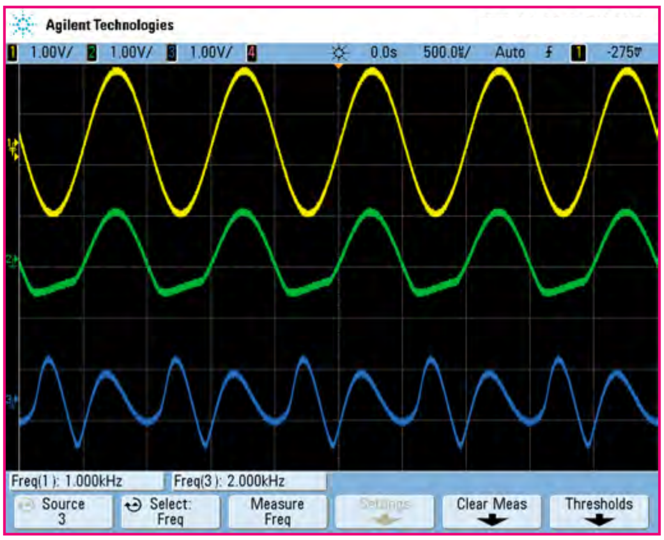


Fig.10: the input signal is still being clipped here, but now we have adjusted VR4 and VR5 to give asymmetrical clipping, resulting in a different type of distortion.

Resistor-Capacitor Substitution Box

with parallel and series RC output

As any engineer, technician or advanced hobbyist will tell you, a resistance substitution box can save a lot of tears and angst. Same comments apply to a capacitance substitution box. Here's one that combines both resistance and capacitance in one box – and you can choose either resistance, capacitance or a combination of both – and that combination can be in series or parallel.

It often seems to be the case that you can never lay your hands on the particular resistor or capacitor you need.

You may be developing a new circuit, repairing an old one, tuning or tweaking equipment, testing test gear... whatever you're doing, circumstances will conspire to ensure that the one component you need is the one that you don't have.

That's when a resistance substitution box or capacitance substitution box can get you out of trouble.

Of course, it's not a permanent 'fix' – it's one that tells you what you need to buy at your next available opportunity.

The beauty of using a true resistance or capacitance substitution box is that the good ones give you a far greater choice of R or C than even discrete components do. So if your circuit needs, say, a 3480Ω resistor, you can provide it.

You can also tell if a 3.3kΩ would do the job or if you need to go to a tighter tolerance. (Incidentally, you can get 3480Ω in the E48 series or above).

In our April 2013 issue, Jim Rowe described a very handy *Resistance Substitution Box*, capable of 'dialling up' any one of a million resistance values between 10Ω and 10MΩ.

Three months later, in July 2012, Nicholas Vinen presented a *Capacitance Substitution Box*, which similarly allowed you to dial up virtually any capacitance between about 30pF and 6µF.

Altronics have taken this concept one step further again, with a combined *Resistance & Capacitance Substution Box*. With a range of 1Ω to 999,999Ω and 100pF to 9.99999µF, it covers the vast majority of resistors and capacitors that you'd normally need in any service, development or troubleshooting work.

Both the resistance and capacitance sections of the box can be used independently via their own pairs of terminals, but can also be connected in series or parallel by means of a 3-position slide switch.

The combined RC network is brought out to another pair of terminals.

The result is a versatile RC box that is more useful than two separate boxes.

It's also smaller than our previous substitution boxes by dint of the use of a pair of six-way, ten-position thumbwheel switches to select the R or C value required.

It's mounted in a sealed ABS enclosure with an overall size of 145 × 105 × 65 (d) mm, with the top-mounted binding posts adding another 16mm.

Residual capacitance

You may be wondering why the minimum capacitance setting in this new box is 100pF when it's easy to get values down to 1pF.

The reason is simple: residual capacitance. When everything is installed on the PCB, even with all care taken to minimise stray capacitance on the PCB, connecting wires, switches and terminals, the residual capacitance is bound to be a lot more than 1pF.

Here, the residual capacitance in the box is about 20pF.

You will need to mentally add this value to any low value of capacitance you select, up to about 500pF; above that, the difference is likely to be swamped by the 10% tolerance of the switched capacitors.

Residual resistance

Similarly, although the lowest selectable resistance value is 1Ω, the residual resistance in the switches, terminals, PCB tracks and interconnecting wiring amounts to about 1.3Ω.

If that sounds a lot, consider that there are six thumbwheel switches,

Decade

By ROSS TESTER

one slide switch and umpteen solder connections to the wiring in the resistance selection and you can see that just a few milliohms in each connection can easily add up to one ohm or more.

So again, when you are selecting low resistance values, you will need to mentally add 1.3Ω to any value below about 100Ω.

Above that value, the 1% tolerance of the switched resistors becomes a dominant factor in the actual resistance value.

The circuit

The full circuit of this *Resistance & Capacitance Substution Box* is shown in Fig.1 overleaf.

It basically consists of six switched banks of resistors and capacitors. The resistance and capacitance sides of the box are independent of each other until specifically connected together by 3-position slide switch S1.

First of all, we'll look at the resistance side. The box works by switching resistors in series. Each switch position adds in another resistor.

Because there are ten positions on each thumbwheel switch, they're called 'decade' switches – they switch in the sequence 1, 2, 3, 4, 5...

So on switch one, position one you'd have 1Ω between the resistance terminals; position two switches in another ohm resistor for 2Ω, position three yet another ohm for 3Ω, and so on.

This is repeated with the other five switches, which in turn, work with 10Ω, 100Ω, 1kΩ, 10kΩ and 100kΩ resistors.

So with all switches in position '9', you would have 9 × 100kΩ (900kΩ) plus 9 × 10kΩ (90kΩ) plus 9 × 1kΩ (9kΩ) plus 9 × 100Ω (900Ω) plus 9 × 10Ω (90Ω) and 9 × 1Ω (9Ω), all in series. Add those all up and you have 999,999Ω (plus the 1.3Ω of residual resistance, of course).

This truth table shows how the binary-coded-decimal switch brings in the capacitors connected to the 1, 2, 4, 8 terminals. Position 5, for example, connects the capacitors on terminals 1 and 4.

DEC	8	4	2	1
0	0	0	0	0
1	0	0	0	1
2	0	0	1	0
3	0	0	1	1
4	0	1	0	0
5	0	1	0	1
6	0	1	1	0
7	0	1	1	1
8	1	0	0	0
9	1	0	0	1

The resistance set by the thumbwheel switches is made available at the top set of red and black terminals.

Capacitance switching

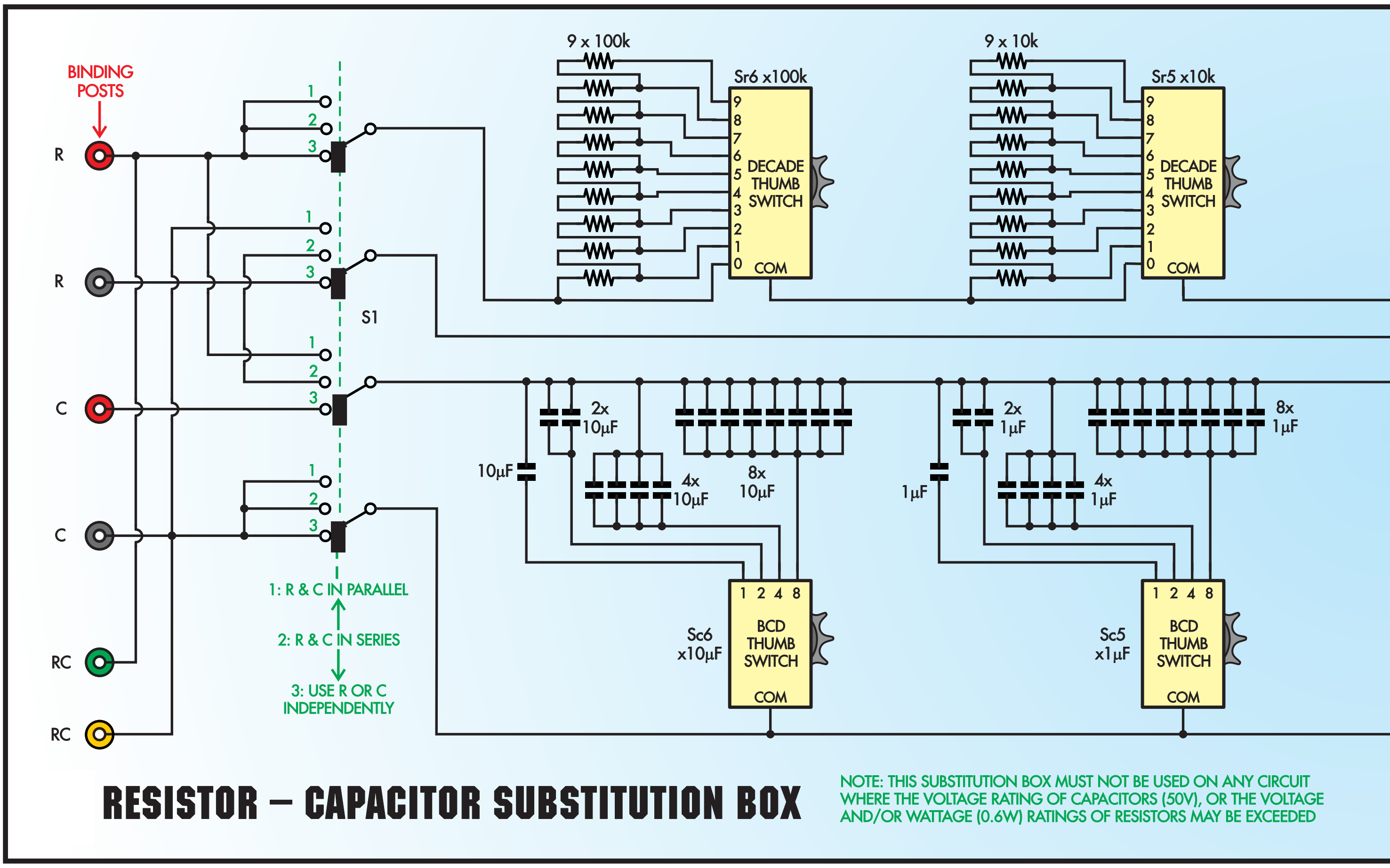
Capacitance selection is done a little differently, using binary-coded decimal (BCD) switches to achieve a similar result with fewer components, saving both space and money (larger capacitors tend to cost more).

And remember that we are switching capacitors in parallel (not series, as with resistors) to obtain larger and larger capacitances.

Connected to the 1, 2, 4 and 8 terminals of the BCD switches are a combination of parallel-connected capacitors.

Looking at the '100pF' switch, a 100pF connects to the '1' terminal, a pair of 100pF (ie, 200pF) connect to the '2' terminal, a 180pF and 220pF (ie, 400pF) connect to the '4' terminal while a 330pF and 470pF (ie, 800pF) connect to the '8' terminal.

Now the BCD coding comes into play. Have a look at the BCD 'truth table' above. In this, '0' means no connection while '1' means a connection. This is all arranged by switch contacts within the BCD switch.



Remember that capacitors in parallel add together, so with the '100pF' switch in positions 1 or 2, you get 100pF and 200pF, respectively. In position 3, the switch connects terminals 1 and 2 together, to give you 300pF. In position 4, you get 400pF, position 5 connects terminals 4 and 1 together to get 500pF, position 6 connects terminals 4 and 2 together (600pF) while 7 connects 4, 2 and 1 together (700pF). Position 8 has only the 800pF connected to it while position 9 connects 8 and 1 to give 900pF.

The second, or x1nF switch, has slightly different values, but they equate to the same thing – 1nF on terminal 1, 2nF on terminal 2, 4nF on terminal 4 and 8nF on terminal 8. Similarly, the third, or x10nF switch, with the 1, 2, 4 and 8 units.

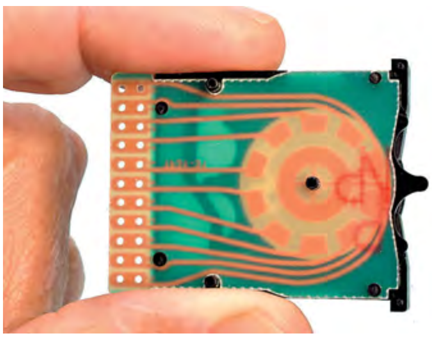
The end result is the same – a maximum of 9.99999µF at the Capacitance (centre) terminals when all capacitance switches are in the '9' position (not forgetting the residual capacitance that we mentioned).

Series/parallel RC

The 3-position slide switch S1 connects the resistance and capacitance sections in series or parallel and the resultant RC network is connected to the third set of terminals, coloured green and yellow to distinguish them from the R and C terminals.

If you're working on a project (or perhaps repairing a device) which uses an RC time constant (such as a timer, frequency generator, filter or even a radio circuit) you can easily

There are two sets of six thumbwheel switches, one set of BCD switches for the capacitors, the other a decade set for the resistors. The six switches click together and are held in position by end plates, as shown here.



Here's how to tell the switches apart: on the decade switch PCB, each switch position has a single track brought out to the rear connector. The BCD switch has a more intricate PCB track pattern.

The six BCD switches (for the capacitors) each have a 9-way header socket attached (only five pins are actually used). The capacitor PCBs plug into these sockets.

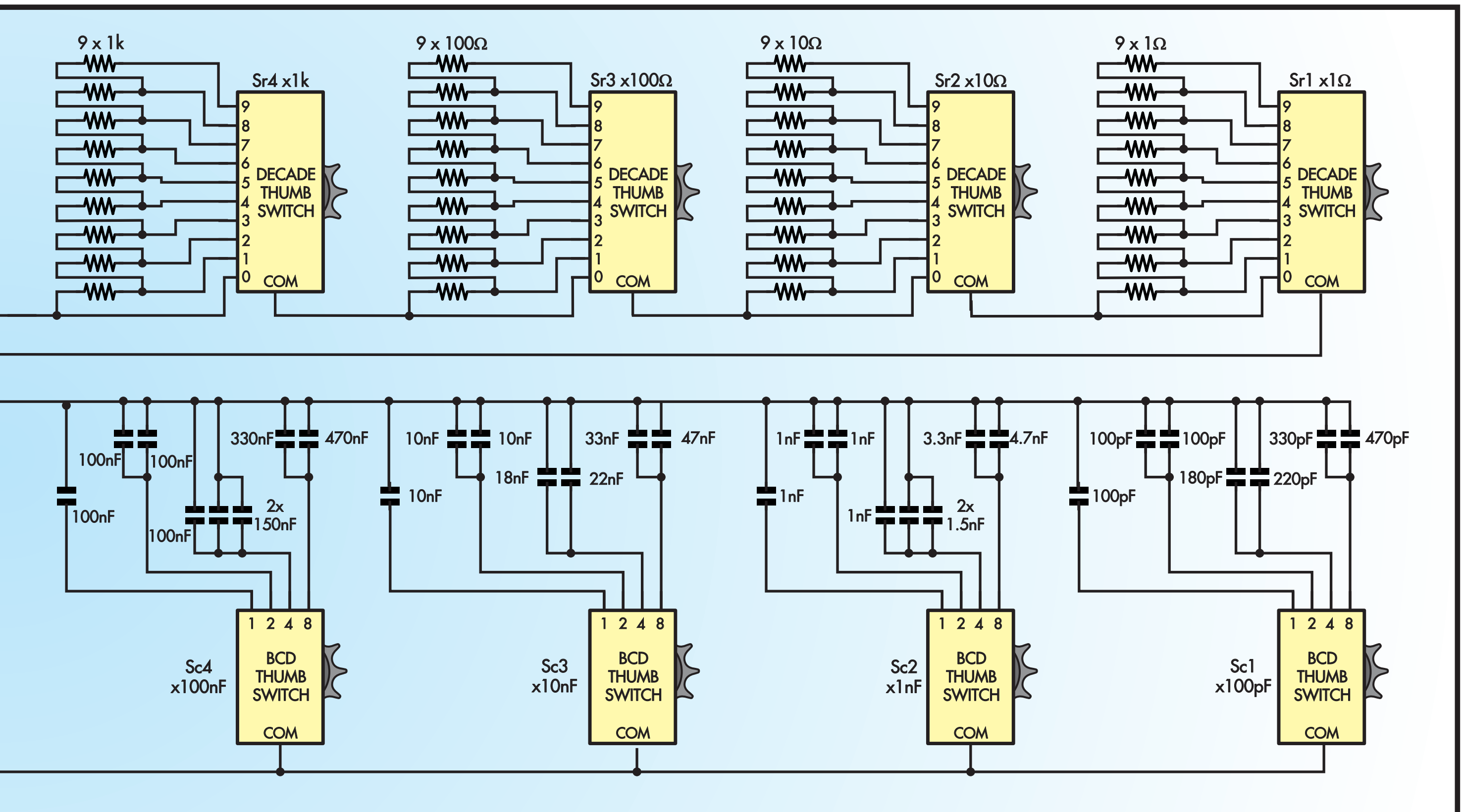


Fig.1: the circuit consists of the various thumbwheel switches bringing resistors and capacitors into circuit. At left, a 3-position slide switch allows series, parallel or independent connection.

achieve this by setting the R and C to their appropriate values and moving the slider switch to either the series or parallel position, depending on the circuit requirements.

Here's where one of the really handy features of this RC box emerges: if the time constant or frequency is not exactly what you're after, it's simply a matter of turning the thumbwheel switches to achieve the desired result.

No more unsoldering and resoldering components – just dial up and go!

When you have got exactly what you need, simply read the values of R and C from the switches, select the same value components and finish/repair/calibrate your project!

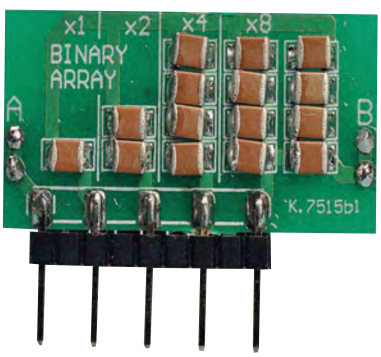
As you can see, an RC box is a pretty handy device to keep on your workbench or service toolbox – and this one is the handiest we've seen.

Construction

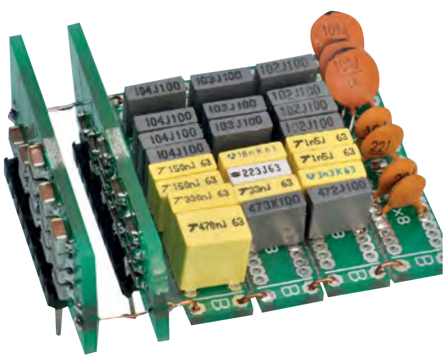
The first step is to assemble the two thumbwheel switch sets. They look similar, so ensure they're not mixed up – the BCD switches have five terminals, and the decade switches have ten.

There are seven small PCBs used in this project, six of which hold the various capacitors and attach to the back of the BCD switch bank.

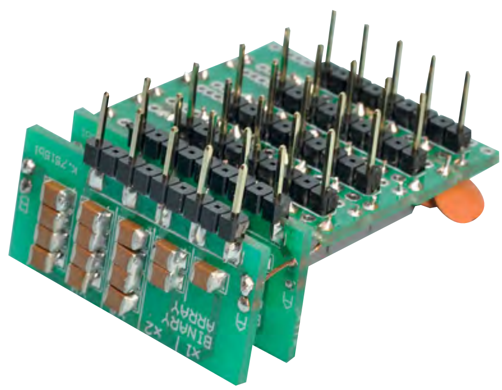
Four of these seven are identical and hold the through-hole capacitors. The other two boards, also identical, hold the 1µF and 10µF capacitors which are all surface-mount devices (SMDs). If you're wondering why SMDs were used



There are two SMD boards which hold the larger value capacitors. All of the capacitors are identical on their respective PCBs.



Four PCBs hold the through-hole capacitors and are mounted side-by-side. Use this photo as a guide to capacitor placement.



And here's the view from the opposite side, showing the six header pin sets underneath, which plug into the BCD thumbwheel switches.

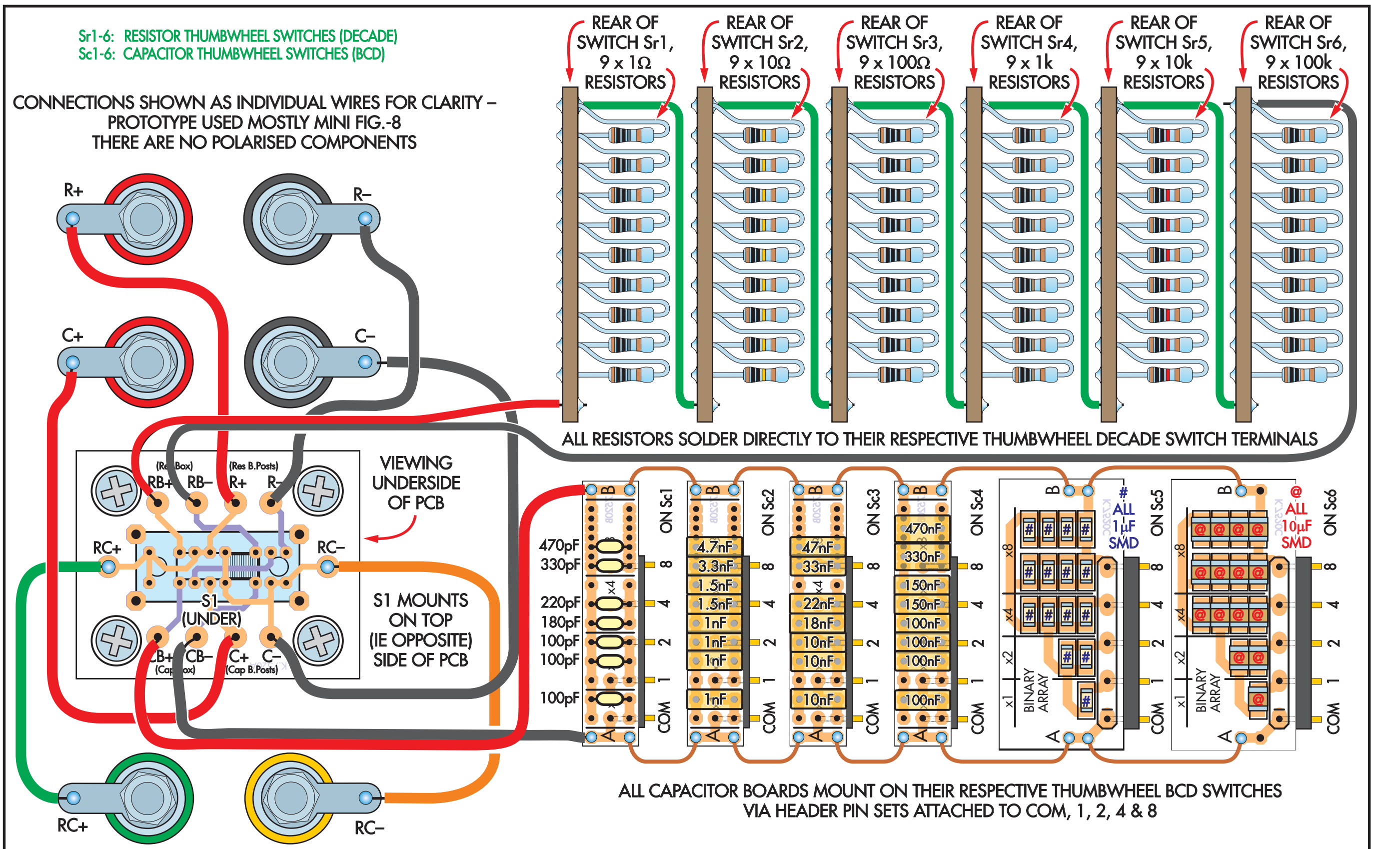


Fig.2: the component layout shows how the resistors and capacitors are mounted – follow this, in conjunction with the photographs, when assembling your *Resistance & Capacitance Substitution Box*.

All resistors mount on the back of the thumbwheel switches in series, with the switches themselves also connected in series, then back to the 3-way switch and output terminals.

on these boards, it's because through-hole versions simply wouldn't fit – apart from the fact that they cost more!

The final board is basically a termination point for the slider switch pins (which mount on it) plus the various flying leads to the other PCBs and to the six terminals.

The resistors (all 54 of them) mount directly to the terminals of the decade switch bank (these terminals are actually small PCBs, but we haven't counted them as they are integral parts of the switches). Nine 1Ω resistors mount on the first switch, nine 10Ω on the second and so on up to the nine 100kΩ on the sixth bank.

This is quite fiddly work – the nine resistors all solder in a tight parallel arrangement, with one lead soldered to the switch contact and its other lead crossing over to the next switch contact. The wrinkle here is that the next resistor in the string also has one lead soldered to the same pad, so you have to ensure that you don't unsolder one as you solder the other!

Our close-up photo at left shows the resistor thumbwheel completely assembled so you can see what we mean.

Once you get the hang of it, it's not that difficult – just tedious. One down, 53 to go. Two down, 52 to go...

These boards are all connected in series: each of the six 'finish' terminals connects, via a short length of hookup wire, to the 'start' terminal on the next switch. The 'start' terminal of switch one and the 'finish' terminal of switch six connect back to the main termination PCB mentioned earlier (and which we'll come to shortly).

Capacitors

As we mentioned earlier, two different types of PCBs hold the capacitors. There are four which secure to the BCD switches 1-4 (100pF, 1nF, 10nF, 100nF) and hold traditional (ie, through hole) capacitors from 100pF to 470nF. The final two boards (1μF and 10μF) are for SMD (surface-mount device) 1μF and 10μF capacitors.

The four boards mount horizontally while the other two (ie, the 10μF and 1μF boards) mount vertically. The main reason that different boards are used for the larger-value capacitors is that through-hole components over 1μF (and especially the 10μF) are too large to mount on the boards so they can fit on the switches.

Once again, assembly isn't too difficult but is complicated by the use of SMDs. Of course, SMDs are used more and more these days (in fact, many components are no longer available in through-hole) so best get used to them!

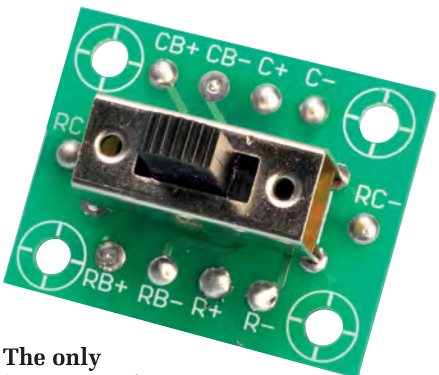
For more detail on the use and soldering of SMDs, refer to the articles on the subject in the July 2010 and February 2014 issues.

Fortunately, all SMDs on each board are identical – there are 15 1µF capacitors on the 1µF switch board and 15 10µF capacitors on the 10µF switch board. Just don't get the 1µF and 10µF types mixed up because they do look similar – note that the 10µF capacitors are somewhat larger. SMD capacitors normally do not come with any markings.

Speaking of mixups, the other four boards are not quite so simple because there is some difference in the component position, not to mention that the component values are all different. Take your time and refer to both the photographs and to the component overlay diagrams.

Unlike the resistance PCBs, all six of the capacitance PCBs connect in parallel – all the 'A' terminals are connected together, as are all the 'B' terminals. The four horizontal boards are connected with short loops of tinned copper wire – the offcuts from the resistor leads are ideal. They should be butted up to each other.

The two vertical-mounting boards have short lengths of tinned copper wire which connect the two boards together (A to A and B to B) and then 'jump across' to join onto the A and B positions on the horizontal boards. The close-up photo will show this more clearly.



The only 'component' on the terminal board is the 3-way switch. All other points connect to the thumbwheels or terminals.

Parts List – Resistor-Capacitor Substitution Box

1 Termination/Switch PCB, Coded K7520A, 28 × 35mm (Altronics)
4 Through-hole capacitor PCBs, Coded K7520B, 35 × 8mm (Altronics)
2 SMD Capacitor PCBs, Coded K7520C, 35 × 16mm (Altronics)
1 ABS Case, 145 × 195 × 65mm, punched and printed (Altronics Cat H0307/K7520)
6 Thumbwheel decade switches (0-9) (Altronics Cat S3302)
6 Thumbwheel BCD switches (0-9) (Altronics Cat S3300)
2 Pairs end caps for thumbwheel switches (Altronics Cat S3305)
1 4-pole, 3-position slider switch (Altronics Cat S2033)
2 40-way pin headers (Altronics Cat P5430)
2 Header pin sockets, 40 pin, 90° (Altronics Cat P5392)
8 Machine screws, M3 × 6mm
4 M3 threaded stand-offs, 12mm
1m hookup wire (or mini fig-8)
Tinned copper wire (if required)
2 short lengths (~50mm) ribbon cable

Capacitors	CODES: µF Value	IEC Code	EIA Code
15 10µF 50V SMD	10µF	10µ	106
15 1µF 50V SMD	1µF	1µ0	105
1 470nF 100V MKT	0.47	470n	474
1 330nF 100V MKT	0.33	330n	334
2 150nF 100V MKT	0.15	150n	154
4 100nF 100V MKT	0.1	100n	104
1 47nF 100V MKT	0.047	47n	473
1 33nF 100V MKT	0.033	33n	333
1 22nF 100V MKT	0.022	22n	223
1 18nF 100V MKT	0.018	18n	183
3 10nF 100V MKT	0.010	10n	103
1 4.7nF 100V MKT	0.0047	4n7	472
1 3.3nF 100V MKT	0.0033	3n3	332
2 1.5nF 100V MKT	0.0015	1n5	152
4 1nF 100V MKT	0.001	1n0	102
1 470pF 50V ceramic	–	470p	471
1 330pF 50V ceramic	–	330p	331
1 220pF 50V ceramic	–	220p	221
1 180pF 50V ceramic	–	180p	181
3 100pF 50V ceramic	–	100p	101

Resistors (1% metal film, 0.6W)
9 100kΩ (Code brown black black orange brown)
9 10kΩ (Code brown black black red brown)
9 1kΩ (Code brown black black brown brown)
9 100Ω (Code brown black black black brown)
9 10Ω (Code brown black black gold brown)
9 1Ω (Code brown black black silver brown)

NOTE: only 1% (5 band) or better resistors should be used for this project to avoid errors.

All six boards 'plug in' to header sockets, which in turn plug in to mating pins on their respective BCD rotary thumbswitches – connecting COM to COM, 1 to 1, 2 to 2, 4 to 4 and 8 to 8.

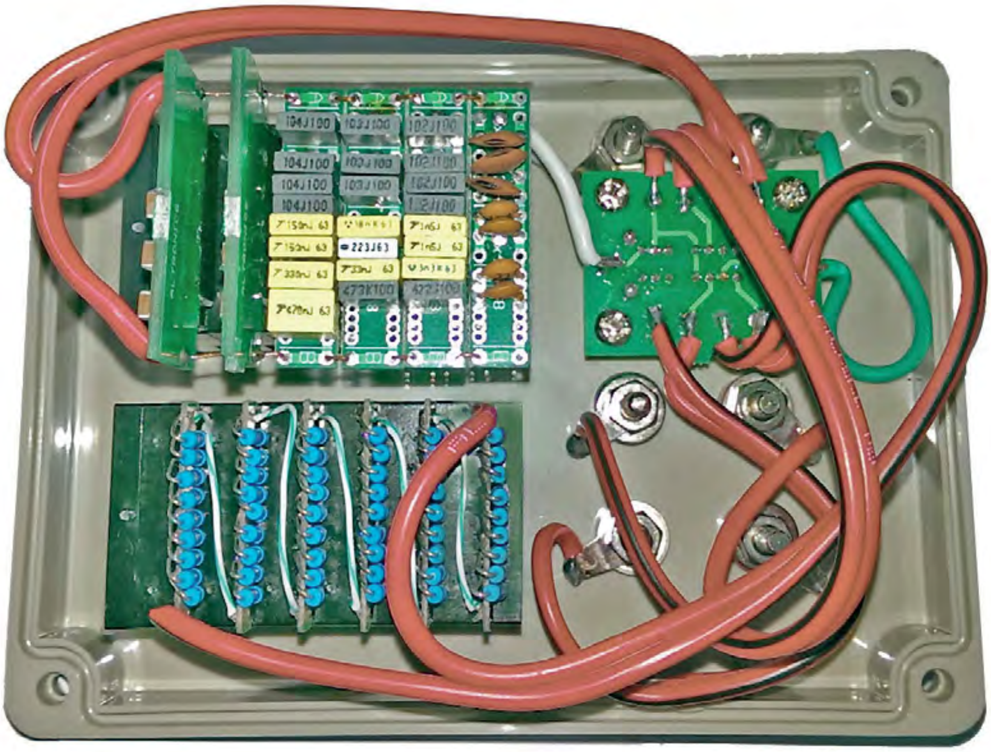
Termination Board

This PCB not only provides an anchor point for the wires coming from the resistance and capacitance board assemblies and going to the six binding posts (terminals), it also provides a mounting point for the two-way, three-position switch which selects between *isolated* R and C, *series* R and C or *parallel* R and C

The switch mounts on the conventional side of the board (it will only go in one way) and the board then mounts upside-down on four 12mm pillars via 6mm M3 screws.

This method enables the switch actuator to poke through the front panel at the right height.

The various wires (ten of them, or five lengths of figure-8) solder to



Finally, here's the completed project, all mounted inside the lid of the case. It has the capacitor switching at top left, resistor switching at lower left, through/parallel/series switch on its PCB at top right and the terminals down the right side.

the exposed copper side of the PCB.

Using the photos as a length guide, cut the wires to appropriate lengths, bare and tin both ends and solder the six solder lugs (which came with the binding posts) to one end. Fit the binding posts to their respective wires.

The opposite ends are now soldered to the PCB – make sure you get the right ones in the right place.

The remaining four wires (or two figure-8s) solder to the 'A' and 'B' positions on the resistance and capacitance boards, as per the layout diagram and photos.

The case

If you're putting this together from the Altronics kit (K7520) it will come with the case already punched and drilled for the thumbwheel switches, parallel/series switch, binding posts and screws – and the top of the case will also be printed, as per our photos.

Checking it out

Give your project the once-over, checking for bad solder joints, misplaced components, etc.

Checking the individual 'R' and 'C' functions is delightfully easy: switch the series/parallel switch to 'off' (ie, fully left) and connect your multimeter on the appropriate range (R or C) to the appropriate substitution box terminals (R or C) and switch through the ranges with the thumbwheels.

Apart from the '000000' settings (or even very low ohms or capacitance), you should find the multimeter reads the same, or at least very close to what your thumbwheels say otherwise, you've got a problem!

If you get no reading at all, it's almost certainly an open circuit/dry joint in your soldering; if you get strange readings, it's more than likely mixed-up components.

As mentoned earlier, with all switches set to zero (on both R and C) it is normal to obtain very low readings – perhaps an ohm or so on resistance and maybe 20pF or so on capacitance. Residual C and R should always be taken into account when working with low settings.

This applies to all R or C substitution boxes, certainly not just this one!

Checking the series or parallel RC combination is not quite so simple – probably the easiest way is to use a moving coil multimeter, set the RC Box to parallel and with your multimeter already connected to the binding posts and on its lowest DC value, switch the RC box to the highest R and C settings.

You should see the voltage rise fairly quickly as the multimeter itself charges the capacitor.

Change the box resistance to a much lower value and the voltage should rise much more quickly. If it does, you can be fairly confident that it's working as it should.

Where from, how much?

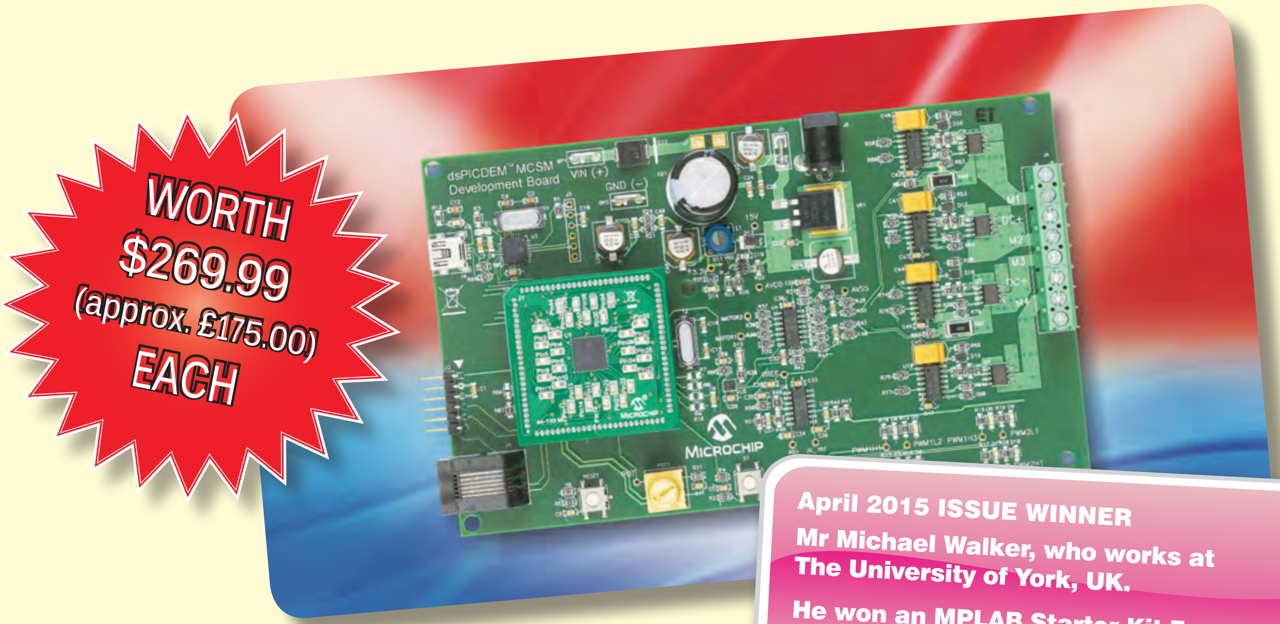
This project was designed by Altronics Distributors, who retain the copyright on the PCBs. Complete kits are available from Altronics via www.altronics.com.au for approx £60 plus p&p. (Catalogue K7520) This includes the pre-printed and punched case.

Win a Microchip dsPICDEM MCSM Development Board

EVERYDAY PRACTICAL ELECTRONICS is offering its readers the chance to win a Microchip dsPICDEM™ MCSM Development Board (DV330021). The development board is targeted to control both unipolar and bipolar stepper motors in open-loop or closed-loop (current control) mode. The hardware is designed in such a way that no hardware changes are necessary for 8-, 6- or 4-wire stepper motors in either bipolar or unipolar configurations. Software to run motors in open-loop or closed-loop with full or variable micro-stepping is provided.

A GUI for controlling step commands, motor parameter input, and operation modes is included. This flexible and cost-effective board can be configured in different ways for use with Microchip's specialised dsPIC33F Motor Control Digital Signal Controllers (DSCs). The dsPICDEM MCSM Development Board offers a mounting option to connect either a 28-pin SOIC device or a generic 100-pin plug-in module (PIM). A dsPIC33FJ32MC204 DSC PIM (MA330017) is included.

The dsPIC DSC devices feature an 8-channel, high-speed PWM with complementary mode output, a programmable ADC trigger on the PWM reload cycle, digital dead-time control, internal shoot-through protection and hardware fault shutdown. These features make the dsPIC DSC an ideal solution for high-performance stepper motor control applications where control of the full-bridge inverter is required.



April 2015 ISSUE WINNER
Mr Michael Walker, who works at The University of York, UK.
He won an MPLAB Starter Kit For Digital Power valued at £85.00.

HOW TO ENTER

For the chance to win a dsPICDEM MCSM Development Board please visit:
www.microchip-comps.com/epe-mcsm and enter your details onto the online entry form.

CLOSING DATE

The closing date for this offer is 30 August 2015.

The latest improved version of our popular electronic thermostat

By JIM ROWE

Here's a new and improved version of our very popular T*empMaster Thermostat*. It's ideal for converting a chest freezer into an energy-efficient fridge, converting a fridge into a wine cooler or controlling heaters in home-brew setups, hatcheries and fish tanks. It controls the fridge/freezer or heater directly via its own power cable, so there's no need to modify its internal wiring. It can even be adapted to control 12V or 24V fridges or freezers.

TempMaster Mk3

Our new *TempMaster* is smaller, easier to adjust, has a wider temperature range and is now virtually immune to relay chatter problems.

The previous version (Mk2) of the *TempMaster* was described in the February 2011 issue of *EPE* and it has been very popular – but, as with most products, actual field use demonstrated that improvements can be made. Some common problems involved 'relay chatter' and motor switch-on/switch-off 'stuttering' when controlling fridges.

Typically, readers also wanted a different temperature range – either above or below the range of 2-19°C we had given the *TempMaster Mk2*.

We had in mind a number of changes and improvements to the 2011 design, but things were brought to a head by a design recently submitted by reader Alan Wilson. He effectively solved the noise sensitivity and relay chattering problem by providing a fast attack/slow decay filtering function, employing the previously unused second comparator in the IC package. So our new version of the *TempMaster* includes his modification.

We have also expanded the temperature adjustment range, reduced the already low quiescent power consumption and it now fits into a smaller and cheaper case. So here is the list of improvements and changes:

- *Much greater noise immunity and hence almost complete freedom from annoying relay chatter and motor switching stutter.*
- *A much wider overall temperature adjustment range (from -23°C to +47°C), which can be set by changing 'max' and 'min' jumper shunts rather than having to change resistor values.*
- *The use of a more efficient low-voltage regulator and CMOS dual op amp, lowering the quiescent power consumption to below 45mW (0.045W) – equating to 1.08Wh/day while running from battery.*

How it works

Fig.1 shows the basic configuration of the *TempMaster Mk3* when it's set up for controlling a fridge or freezer.

The heart of the circuit is the remotely-mounted LM335Z temperature sensor, TS1. The LM335Z acts similarly to a special kind of zener diode, but its voltage drop varies in direct proportion to absolute temperature, having a value of 0V at 0 kelvin (–273°C) and rising linearly by 10mV for every kelvin (or °C) rise in temperature.

This is shown in the graph of Fig.2. At a temperature of –10°C (263K), the voltage drop of the LM335Z is very close to 2.63V. Similarly, at 40°C (313K), it rises to 3.13V. We use this change in voltage to control the temperature of our fridge/freezer or heater by comparing the sensor's voltage with a preset reference voltage.

The comparison is made by IC1a, one section of an LMC6482AIN dual CMOS op amp which is connected as a comparator. For cooling control, the sensor voltage V_{SENSOR} is fed to the non-inverting input, pin 3, of IC1a via a 1.2kΩ resistor, while the reference voltage V_{REF} is taken from adjustment trimpot VR1 and fed to the inverting input, pin 2.

If V_{SENSOR} is lower than V_{REF} (because the temperature of TS1 is lower than that corresponding to V_{REF}), the output of IC1a will be low – close to 0V. But if the temperature being sensed by TS1 should increase to the set threshold, V_{SENSOR} will rise just above V_{REF} and the output of IC1a will switch high – to almost +12V.

Heating

The reverse sequence of events happens when the circuit is configured for heating control rather than cooling. In this mode, sensor TS1's voltage V_{SENSOR} is fed to the inverting input of IC1a, while the reference voltage V_{REF} is fed to IC1a's non-inverting input via the 1.2kΩ resistor. (In other words, the two voltages are swapped around.)

As a result the output of IC1a remains low when V_{SENSOR} is higher than V_{REF} – but, switches high as soon as V_{SENSOR} falls below V_{REF}.

Hysteresis

Returning to the cooling control configuration shown in Fig.1, note the 10MΩ resistor connected between the output of IC1a (pin 1) and its non-inverting input (pin 3). This is to provide a very small amount of positive feedback.

We do this so that once pin 1 has switched high, the actual voltage fed to pin 3 will be slightly higher than the sensor voltage V_{SENSOR} (about 1mV higher, in fact). As a result, V_{SENSOR} needs to fall slightly below V_{REF} before the voltage at pin 3 drops to the level matching V_{REF}. But then pin 1 suddenly switches low again, which causes the voltage at pin 3 to drop back to V_{SENSOR}.

So the effect of this small amount of positive feedback is to create a small difference between the comparator's turn-on and turn-off voltage levels (and the corresponding temperatures).

This is called 'hysteresis' and is designed to minimise any tendency for the comparator to oscillate or 'stutter' at the switching thresholds – especially the turn-off threshold.

Now we come to the improvement proposed by reader Alan Wilson, involving diodes D3, D4 and IC1b. Together with the 10µF capacitor and the second 10MΩ resistor, D3 and D4 form a fast-attack/slow-decay filter. This works in conjunction with IC1b (connected as a comparator) to ensure that transistor Q1 and the power switching relay are able to turn on quite rapidly as soon as the output of IC1a switches high, but cannot switch off again for 30 seconds or so after the output of IC1a has dropped low.

This is because the 10µF capacitor can charge up quickly via D3, but can only discharge quite slowly via D4 and the 10MΩ resistor – and only when the output of IC1a has dropped low, in any case.

IC1b also has a modest level of positive feedback applied, via the 220kΩ resistor linking pins 7 and 5. This also helps ensure that there can be no relay stuttering during either turn-off or turn-on.

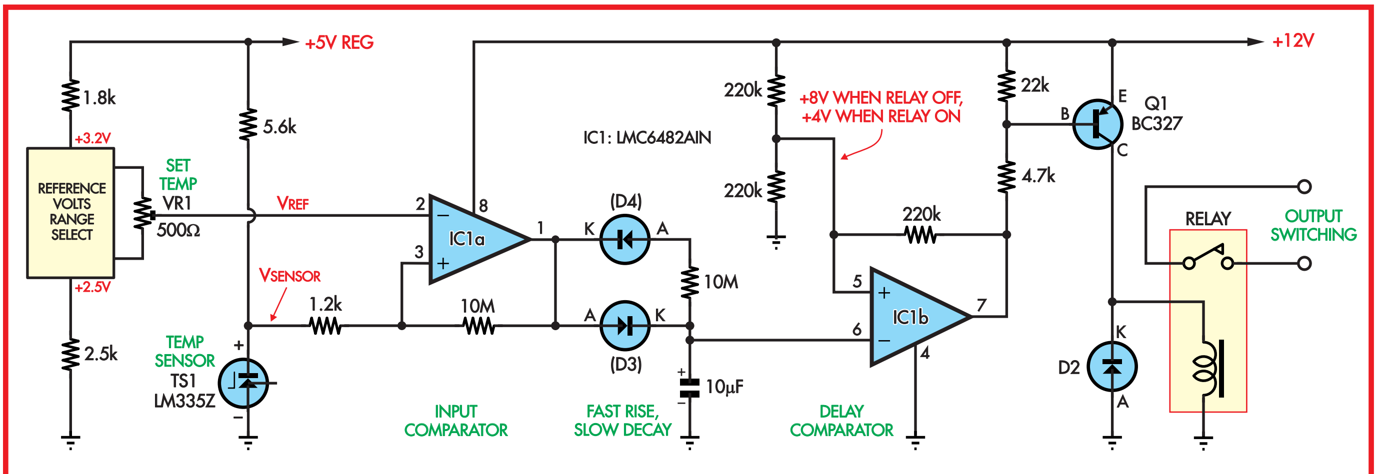


Fig.1: this simplified circuit shows the basic operation. The full circuit is shown overleaf in Fig.3.

Fig.2: the LM335Z sensor voltage changes with its temperature, and that change is linear from way below zero up to above the boiling point of water. Actual operating range is -40°C to +100°C.

The full circuit

Now let's look at the full circuit of Fig.3 to consider the finer points of operation. Temperature sensor TS1 plugs into socket CON2, which connects to test point TP2 and one end pin of links LK1 and LK2. It also connects to the regulated +5.0V rail via a 5.6kΩ resistor, which feeds the sensor a small bias current. The regulated +5.0V rail is provided by REG1, an LP2950ACZ device.

The reference voltage to be compared with the sensor voltage is derived from the same regulated +5.0V supply rail, via a voltage divider formed by the 1.8kΩ resistor (at the top) – plus a string of 200Ω and 100Ω resistors and finally the 2.4kΩ resistor at the bottom.

The divider provides a set of five different tapping voltages, with +3.2V available at the top and +2.5V at the bottom. Link set LK3 allows you to select one of three voltage levels as the temperature range maximum, while link set LK4 allows you to select one of another three voltages as the temp range minimum.

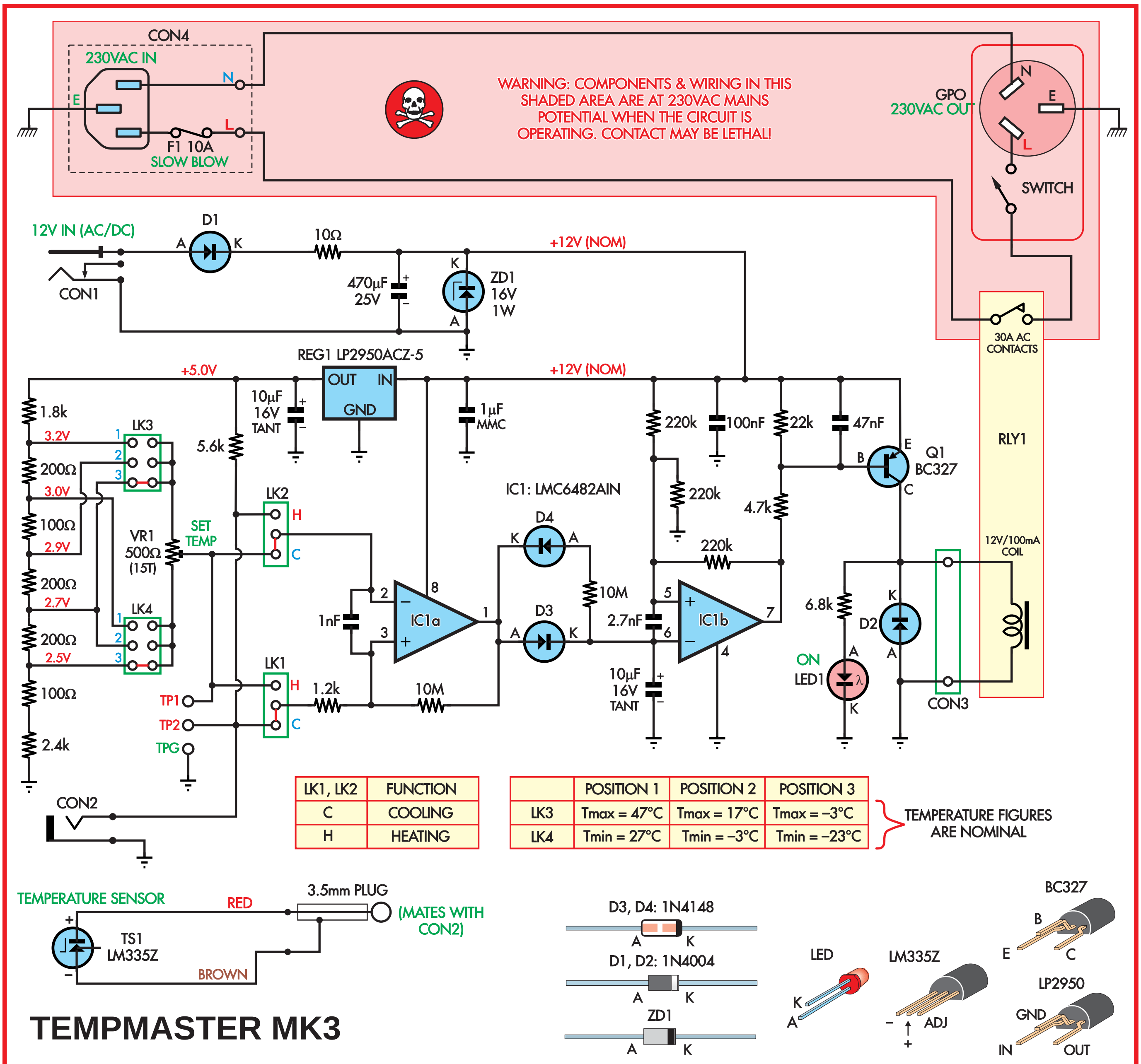
The temperature setting 'fine tuning' is done using VR1, a 500Ω multi-turn trimpot. Its two ends are connected to LK3 and LK4 respectively, so whichever maximum and minimum temperatures have been selected using these links, VR1 then allows you to select any specific V_{REF} in this range, corresponding to your desired threshold or 'set point' temperature.

For example, if you have set LK3 to position 3 to give a maximum V_{REF} of 2.7V, and have also set LK4 to position 3 to give a minimum V_{REF} of 2.5V, VR1 will then let you select any voltage between these two limits. This means you'll be able to select any threshold temperature between about –3°C and –23°C. Get the idea?

Note that the selected reference voltage V_{REF} is made available at test point TP1, while the sensor voltage V_{SENSOR} is always available at TP2.

These two voltages go to links LK2 and LK1, which are used to select either the heating (H) or cooling (C) mode of operation. As mentioned earlier, this involves simply swapping which of the two voltages, V_{REF} and V_{SENSOR}, is passed to the non-inverting input of IC1a, and which is fed to the inverting input.

The rest of the circuit is very similar to the basic outline in Fig.1. The only real differences are the addition of small filter capacitors between both inputs of IC1a and IC1b (to improve noise immunity), and the addition of LED1 with its 6.8kΩ series resistor, across the relay coil. This is to provide an indication of when the relay is energised.



LK1, LK2	FUNCTION
C	COOLING
H	HEATING

	POSITION 1	POSITION 2	POSITION 3
LK3	Tmax = 47°C	Tmax = 17°C	Tmax = –3°C
LK4	Tmin = 27°C	Tmin = –3°C	Tmin = –23°C

Fig.3: this full circuit of our new *TempMaster Mk3* has many similarities with the simplified version of Fig.1. While the control circuitry operates from low voltage and is isolated, it is switching mains so must be regarded as dangerous.

All of the circuit operates from 12V DC fed via CON1, polarity protection diode D1 and a 10Ω resistor, which limits the current through zener diode ZD1 if the voltage rises above 16V.

The supply can come from a 12V plugpack or battery, and since the current drain is only around 100mA when the relay is switched on and less than 4mA when it's off, only a small battery or plugpack is required. This should make the *TempMaster Mk3* very suitable for use with solar power systems.

Construction

Nearly all of the components used in the *TempMaster Mk3* circuit are mounted on a PCB available from the *EPE PCB Service*, measuring 104 × 80mm and coded 21108141. The board has rounded cut-outs in each corner, so it fits inside a sealed polycarbonate case measuring 115 × 90 × 55mm, sitting on the tapped pillars moulded into the bottom of the case.

We have used a rugged 12V relay (RLY1) rated to switch 250VAC at up to 30A so that it can easily handle typical fridge, freezer or heater loads. The connectors for the 12V DC input (CON1) and remote temperature sensor TS1 (CON2) are mounted on the right-hand side of the board, accessed via matching holes on that side of the case. The 'set temperature' trimpot VR1 is mounted between these two connectors and is also accessed by a small hole, while the 'relay on' indicator LED1 is visible via a similar small hole below CON2.

The only components not mounted on the PCB inside the *TempMaster Mk3* itself are the fused IEC mains input connector (CON4) and the switched 3-pin mains outlet or

GPO. The latter is mounted on the lid, while the former mounts in the left-hand side of the case (in a matching cut-out).

Note that CON4 should be fastened inside the case using two 10mm nylon screws and nylon hex nuts.

When wiring the board, follow the internal photos and Fig.5 closely.

Begin wiring up the board by fitting the three terminal pins (used to provide test points TP1, TP2 and TPG). These go at centre right on the board. Then fit DC input connector CON1, temperature sensor socket CON2 and the two-way terminal block CON3 (used for the relay coil wires). If you want to use a socket for IC1 this can be fitted now as well.

You can also mount the two three-way SIL headers for LK1 and LK2, which are located just to the left of TP1. Then fit the two 3x2 DIL headers for LK3 and LK4, which go just above LK2.

Next, install the various fixed resistors, making sure each one goes in its correct position. Check their values with a DMM just before it's fitted to the board. Then fit trimpot VR1, between CON1 and CON2.

The five non-polarised polyester and MMC capacitors can go in next, followed by the two 10µF tantalums and finally the 470µF electrolytic. Note that the last three are polarised and must go in the correct way around.

Then fit diodes D1-D4, zener diode ZD1 and transistor Q1, again paying attention to polarity. LED1 should be mounted vertically and with the bottom of its body about 15mm above the board (the leads will be bent by 90° later). Make sure the LED is oriented so that its 'flat' is near the top of the board and its longer anode lead is passing through the lower hole in the board.

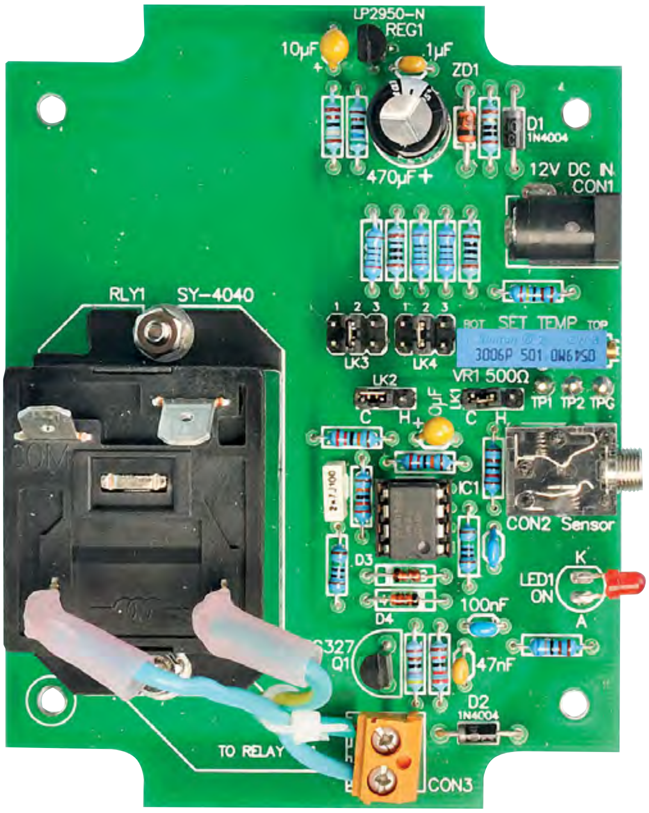
Then solder REG1, followed by IC1 – soldering it in place if you're not using an IC socket.

Relay RLY1 is attached to the board using two M4 × 10mm machine screws, with flat washers, lockwashers and hex nuts. Before you mount it, you need to cut a small piece from the relay's mounting flange at the switching contacts end, as shown in Fig.5. (This is to provide clearance for the body of CON4, when it's fitted later.) The soft plastic can be cut quite easily using a small hacksaw and the cut edges smoothed using a small file.

Then mount the relay on the PCB with its coil connection spade terminals at the bottom and its contact connectors at the top, again as shown in Fig.5. Also make sure that you fit the relay mounting screws facing upwards – that is, with their heads under the board and the nuts and washers above the relay-mounting flanges. Otherwise, the PCB assembly won't fit properly down inside the case.

With the PCB now complete, you drill and cut the various holes needed in the case and its lid. The drilling and cutting details are shown in Fig.7.

Note that the cut-out in the rear long side of the case/box for fused IEC mains inlet CON4 extends almost to the very top – but not quite. Drill and file the cut-out first so



Full-size photo of the assembled PCB. All components (with the exception of the IEC mains input socket and the GPO) mount on this board. Note the double-insulating layer of heatshrink tubing over the coil wiring between the PCB and the coil spade terminals.

that it extends almost to the top of the outer box side and then carefully extend the top using a small file, until CON4 just slips inside.

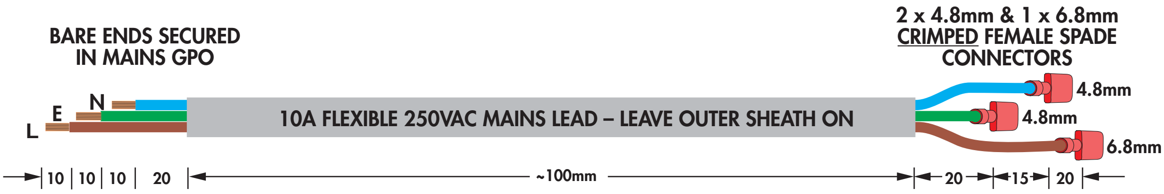
Once the case is prepared, lower the PCB assembly down into the main part of the case until it's resting on the standoff pillars. Then decide where the leads of LED1 will need to be bent outward by 90°, so it will just protrude from the matching hole in the side of the case. When you have bent the LED leads to achieve this, lower the PCB assembly into the case again and screw it into place using four M3 × 6mm machine screws, which mate with the metal nuts moulded into the standoffs in the bottom of the case.

Then fit the IEC mains input connector CON4 into its cut-out, and secure it with two M3 × 10mm nylon screws and nuts.

Mount the mains outlet GPO on the case lid, with its 'rear side' passing through the matching rectangular cut-out. This is done by unclipping the outer dress cover plate, to reveal the various recessed mounting holes which are provided.

The holes you'll be using here are those that are spaced 84mm apart, along the 'east-west' centreline of the GPO. You need to attach the GPO to the case lid using a pair of M4 × 15mm pan-head screws passing down through these holes and fitted with star lockwashers and M4 nuts inside.

Fig.4: the cable connecting the input and output sockets should be cut from a 10A 3-core mains cable offcut.



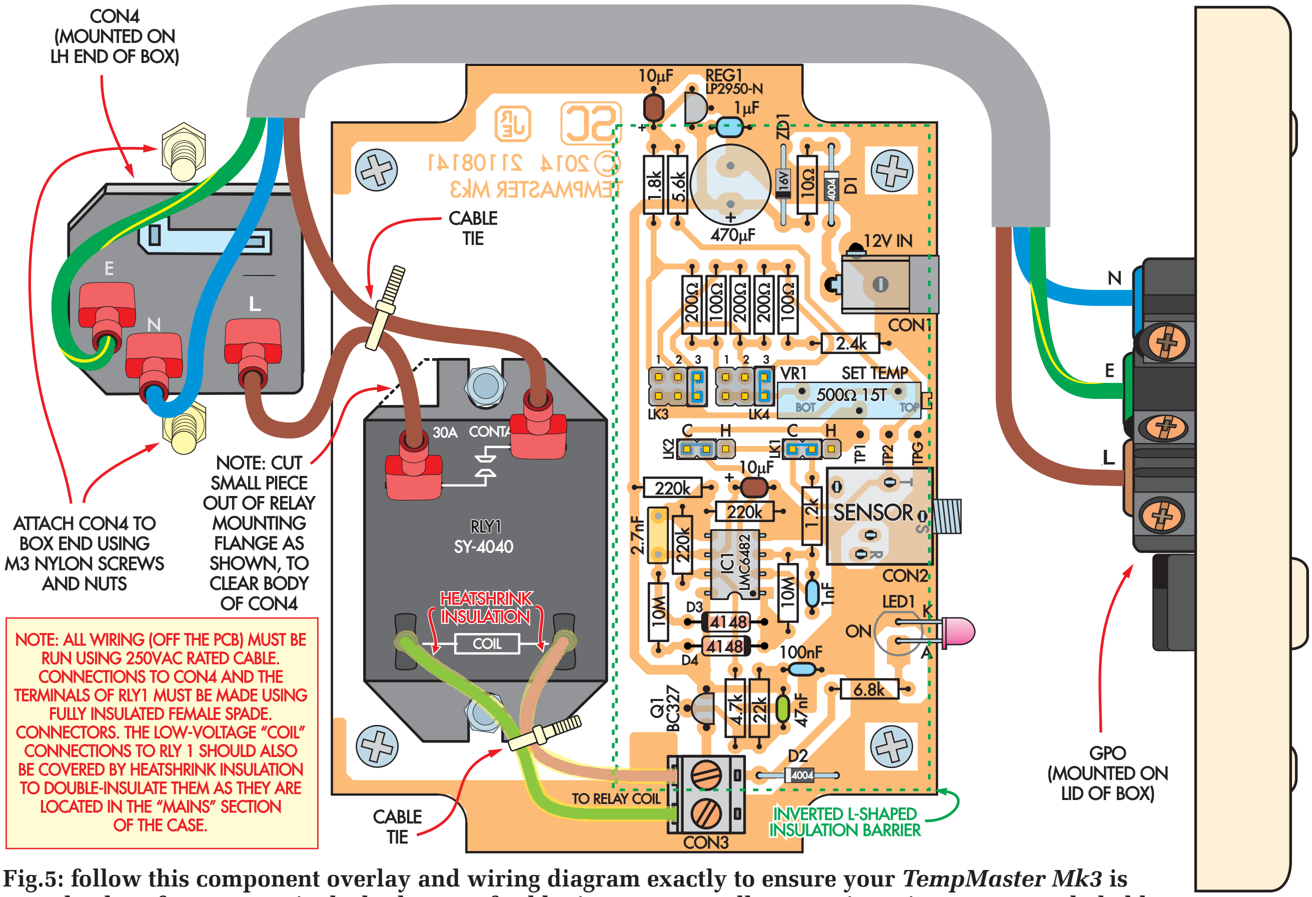


Fig.5: follow this component overlay and wiring diagram exactly to ensure your *TempMaster Mk3* is completely safe. Note particularly the use of cable ties to ensure all connecting wires are securely held – that's also the reason we use a piece of flexible 10A mains cable with its outer sheath left in place as much as possible.

Tighten these up firmly to make sure that the GPO can't work loose.

Don't fit the GPO's dress cover plate at this stage. It's clipped on later – after the lid is finally screwed onto the case, because the cover plate just interferes with the lid-to-case assembly screw heads.

Next you need to prepare the mains connection cables which link the GPO to the IEC mains connector and the contacts of RLY1.

Fig.4 shows a same-size diagram of the mains connecting cable. It makes sense to use a length of thin mains-rated 10A flex for this as you will not only obtain the insulation level required, but leaving the outer sheath on the cable also keeps the conductors together.

Note that the blue (neutral) and green/yellow (earth) wires from the GPO have 4.8mm fully insulated female spade connectors crimped firmly to their 'far ends', while the brown (live) wire has a 6.8mm spade connector attached. The shorter brown (live) wire connecting from the IEC connector live to the relay switch contact also has insulated spade connectors at both ends, one 4.8mm and one 6.8mm wide.

Make sure you attach all of these spade connectors very firmly using a rachet-type crimp connector, so they will give reliable long-term connections.

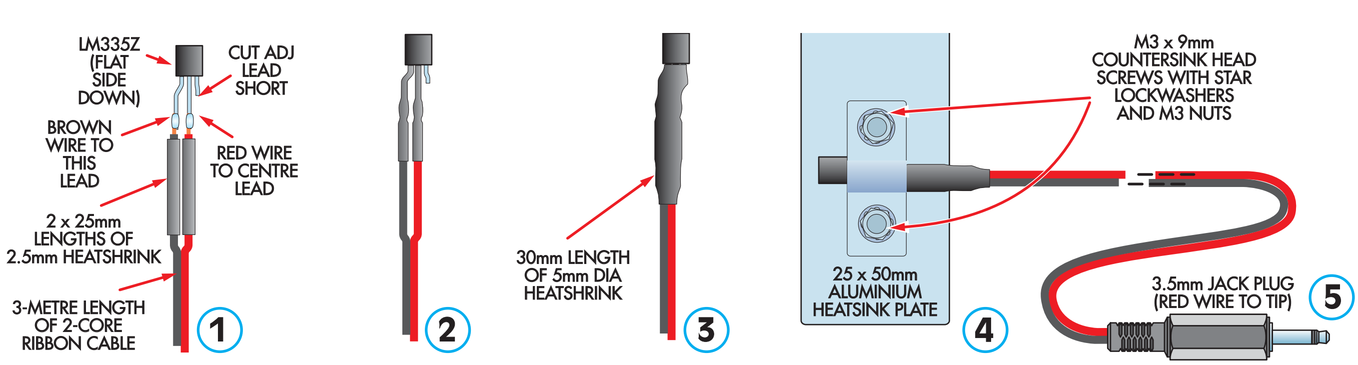


Fig.6: steps in wiring the LM335Z temperature sensor. In step 1, the unwanted 'ADJ' lead is cut off, two wires are soldered to the other pins and then covered with heatshrink. In step 2, the heatshrink is slid up and over the soldered leads and shrunk, followed by a larger length of heatshrink over the whole assembly. In step 4, you secure the sensor to a heatsink, then finally in step 5 connect the two wires to a 3.5mm jack plug.

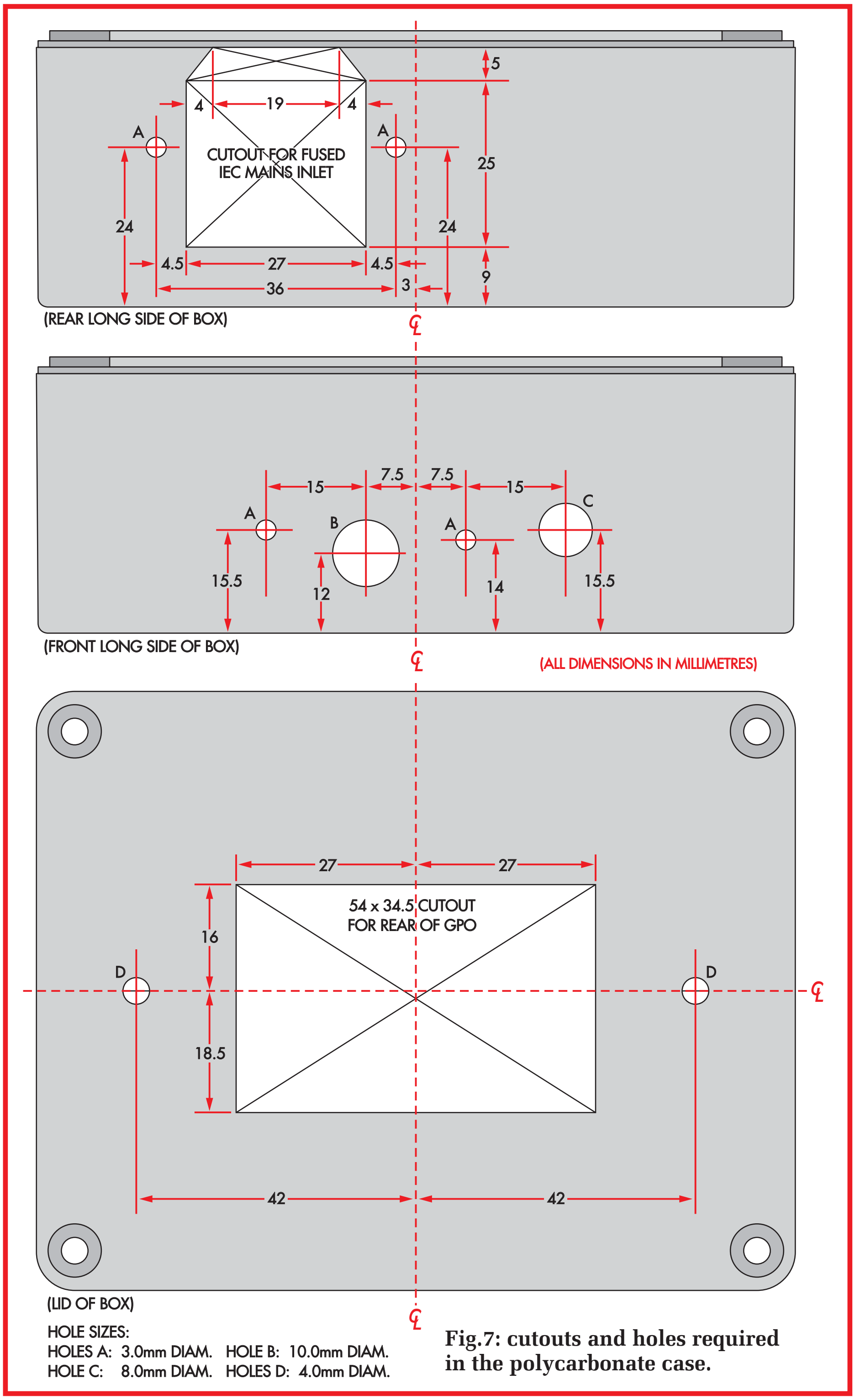


Fig.7: cutouts and holes required in the polycarbonate case.

Last, you can make up the two short wires which are used to connect the coil of RLY1 to terminal block CON3. These can be made up from medium-duty insulated hookup wire, with each one having a 4.8mm insulated female spade connector crimped to one end.

Once all the wires have been prepared, you can use them all to connect everything up as shown in Fig.5. This will complete the wiring of the *TempMaster Mk3*, but before you screw on the lid of the case to finish assembly, fit a nylon cable tie to the mains wiring as shown in Fig.5 and the internal photo. This is to ensure that should any of the spade connectors somehow work loose, there is no way that it can swing around and make contact with any of the low-voltage wiring.

You can also fit another cable tie around the wires from the relay coil to CON3, to make sure these will also hold each other in place.

Now you can fit jumper shunts to the two 3-way SIL header strips LK1 and LK2, in the centre of the PCB, depending on whether you're going to be using the *TempMaster Mk3* to control cooling or heating. You should also fit jumper shunts to one of the three positions on both DIL header strips LK3 and LK4, to set the maximum and minimum of the temperature adjustment range you wish to use.

Safety insulation

Because there are low-voltage components in close proximity to the mains outlet when the case is closed, it is essential to make sure they can never come in contact with each other.

We do this with an insulating barrier, cut from a piece of Presspahn, Elephantide or similar insulation and bent it into an 'L'-shape (as shown in Fig.8). This slides down the edge of the relay, keeping the mains and low voltage sides separate.

A dollop of glue on the edge of the relay and the surface of the PCB alongside will hold the barrier in place when the top goes on.

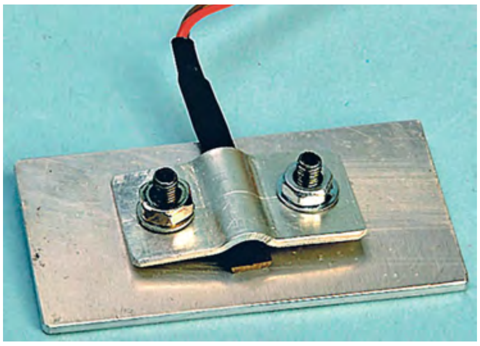
Fit the rubber sealing strip around the groove in the underside of the case lid and then screw the lid to the case using the four screws provided. Then you'll be able to clip the cover plate back on the GPO, to complete the assembly of the *TempMaster Mk3* itself.

Making the remote sensor

The details for the temperature sensor are shown Fig.6. The first step is to cut short the unwanted third lead of the LM335Z sensor and then solder the ends of a 2-core ribbon cable to the other two leads after slipping 25mm lengths of 2.5mm-diameter heatshrink sleeving over each one.

After the solder cools, the sleeves are then moved up until they butt hard against the body of the LM335Z. Then they are heated to shrink them in place (step 2). Then a 30mm length of 5mm-diameter heatshrink sleeving is slipped along the cable and over the other sleeves, and heated in turn to shrink it in place as well (step 3).

Prepare the sensor's heatsink assembly by drilling two 3.5mm diameter holes on the centre line of the 50 × 25mm aluminium plate, 18mm apart.



A close-up of the heatsink and clamp assembly for the LM335Z temperature sensor.

The bottom of both holes should be countersunk to accept countersink-head screws passed up from underneath.

Next, make the 30 × 10mm piece of 1mm aluminium into a clamp piece, by bending its central 8mm section into a half-round shape to fit snugly over the LM335Z body. After this drill, 3.5mm holes in the flat ends of this clamp piece, 18mm apart again to match the holes in the larger plate. You should then be able to assemble the probe with the LM335Z clamped to the top of the plate 'flat side down', and the screws tightened down using M3 nuts and star lockwashers (step 4).

Complete the sensor assembly by fitting the 3.5mm mono jack plug to the other end of the two-core ribbon cable, connecting the red wire to the 'tip' lug and the brown wire to the 'sleeve' lug (step 5).

Initial checks

Before doing anything else, use your multimeter or DMM (set to a low-ohms range) to check between the earth pin of the IEC connector (CON4) and the earth outlet of the GPO. You should get a reading of zero ohms here (this checks the integrity of the earth connection).

Then fit a 10A slow-blow M205 fuse into the fuseholder in the IEC connector. Do not connect 230VAC power to the unit until you have done the set-up adjustments. All setup is done using the low-voltage supply only.

DO NOT CONNECT 230VAC power without the lid in place, to eliminate the risk of electric shock.

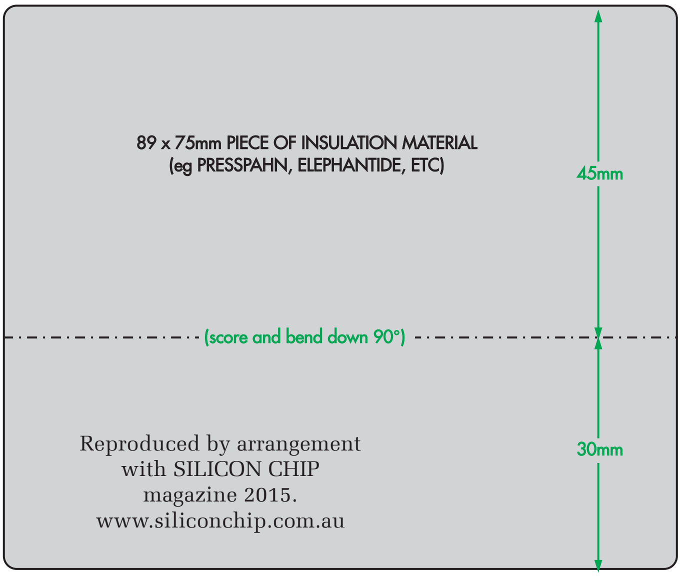


Fig.8: L-shaped insulation barrier inserted between the low voltage components and the mains wiring.

Parts List – TempMaster Mk3

1 Polycarbonate case, light grey, 115 × 90 × 55mm
1 PCB, available from the *EPE PCB Service*, code 21108141, 80 × 104mm
1 SPST relay, 30A contacts with 12V/100mA coil
1 2.1mm or 2.5mm concentric DC connector, PC-mounting, to suit plugpack (CON1)
1 3.5mm switched stereo socket, PC-mounting (CON2)
1 2-way terminal block, PC-mounting (CON3)
2 3-pin SIL header strip, PC-mounting (LK1, LK2)
2 3×2-pin DIL header strip, PC-mounting (LK3, LK4)
4 Jumper shunts
3 1mm-diameter PCB terminal pins
1 IEC panel-mount mains socket with fuse (CON4)
1 Single 250VAC switched general purpose outlet (GPO)
1 10A M205 fuse cartridge, slow blow
1 105 × 75mm piece Presspahn insulation
4 M3 6mm machine screws, pan head
2 M4 10mm machine screws, pan head
2 M4 15mm machine screws, pan head
4 M4 hex nuts with flat and lockwashers
2 M3 10mm Nylon screws, pan head, with Nylon hex nuts
1 205mm length of 10A 3-core mains flex
1 60mm length of 10A brown mains wire
2 70mm lengths of medium duty insulated hookup wire
6 Nylon cable ties
2 6.8mm insulated female spade connectors for 1.2mm wire
5 4.8mm insulated female spade connectors for 1mm wire
1 3m length of 2-conductor ribbon cable
1 25 × 50 × 3mm aluminium sheet
1 30 × 10 × 1mm aluminium sheet
2 25mm lengths of 2.5mm heatshrink sleeving
1 30mm length of 5.0mm heatshrink sleeving
2 M3 9mm machine screws, countersink head
2 M3 hex nuts & star lockwashers
1 3.5mm mono jack plug

Semiconductors

1 LMC6482AIN dual CMOS op amp (IC1)
1 LP2950ACZ-5 micropower LDO regulator (REG1)
1 LM335Z temperature sensor (TS1)
1 BC327 PNP transistor (Q1)
1 16V 1W zener diode (ZD1)
1 3mm red LED (LED1)
2 1N4004 1A diodes (D1,D2)
2 1N4148 signal diodes (D3,D4)

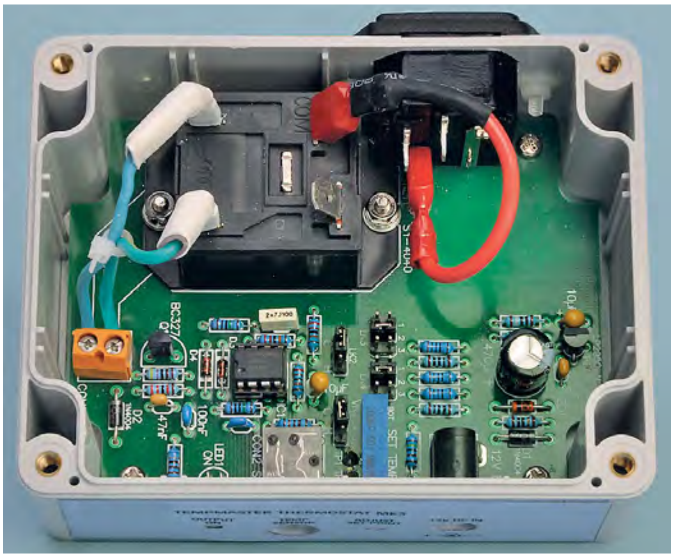
Capacitors

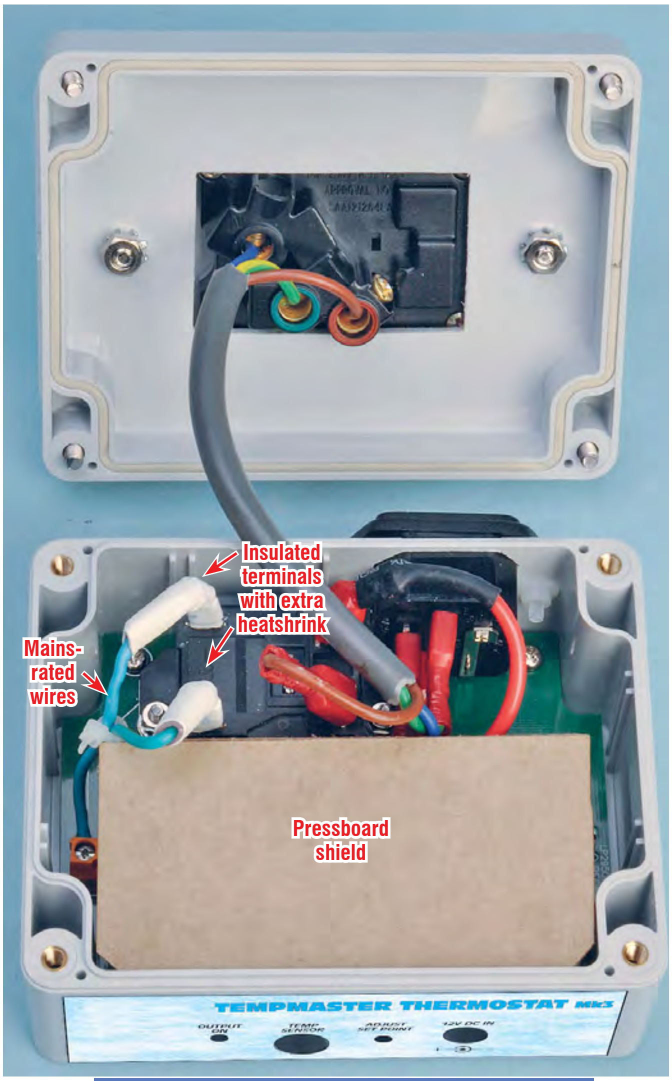
1 470µF 25V RB electrolytic
2 10µF 16V tag tantalum
1 1µF monolithic multilayer ceramic
1 100nF monolithic multilayer ceramic
1 47nF MKT or ceramic/MMC
1 2.7nF MKT or ceramic/MMC
1 1nF MKT or ceramic/MMC

JAYCAR ELECTRONICS have released a 'short form' kit for the TempMaster Mk3
It includes a PCB with relay and onboard components, plus temperature sensor and mounting plate.
Cat KC-5529
Approx £20+p&p

Resistors (0.25W 1% unless specified)

2 10MΩ 3 220kΩ 1 22kΩ 1 6.8kΩ 1 5.6kΩ 1 4.7kΩ
1 2.4kΩ 1 1.8kΩ 1 1.2kΩ 3 200Ω 2 100Ω
1 10Ω 0.5W 5%
1 500Ω horizontal 10-turn cermet trimpot (VR1)





Internal views of the *TempMaster Mk3* – above, with the PCB in place and at right, fully assembled with shield.

Setting it up

This is done by adjusting trimpot VR1 (using a small screwdriver through the access hole in the front panel) to produce the reference voltage level at test point TP1 corresponding to the average temperature you want the *TempMaster Mk3* to maintain.

First plug the 12V DC cable from your plug pack or battery supply into CON1 at the right-hand end of the box – do not plug the mains supply in yet. Then use your DMM to measure the DC voltage between TP1 and TPG.

The voltage should be somewhere between the maximum and minimum levels you have set using the links of LK3 and LK4. Select the temperature you want from the horizontal axis of the graph in Fig.2, and adjust VR1 to obtain the corresponding DC value on the vertical axis.

All that remains now is to mount the remote sensor inside the fridge or freezer cabinet, or inside the hothouse or seed-germinating cabinet, attaching the sensor's heatsink plate to the side of the cabinet using two short lengths of 'gaffer' tape.

Then you can run its ribbon cable outside, holding it down with further strips of gaffer tape so it will pass neatly under the rubber door seal when the door is closed.

Capacitor Codes

	No	Value	μF Value	IEC Code	EIA Code
❑	1	1μF	1μF	1000n	105
❑	1	100nF	0.1μF	100n	104
❑	1	47nF	0.047μF	47n	473
❑	1	2.7nF	0.0027μF	2n7	272
❑	1	1nF	0.001μF	1n	102

Resistor Colour Codes

	No.	Value	4-Band Code (1%)	5-Band Code (1%)
❑	2	10MΩ	brown black blue brown	brown black black green brown
❑	3	220kΩ	red red yellow brown	red red black orange brown
❑	1	22kΩ	red red orange brown	red red black red brown
❑	1	6.8kΩ	blue grey red brown	blue grey black brown brown
❑	1	5.6kΩ	green blue red brown	green blue black brown brown
❑	1	4.7kΩ	yellow violet red brown	yellow violet black brown brown
❑	1	2.4kΩ	red yellow red brown	red yellow black brown brown
❑	1	1.8kΩ	brown grey red brown	brown grey black brown brown
❑	1	1.2kΩ	brown red red brown	brown red black brown brown
❑	3	200Ω	red black brown brown	red black black black brown
❑	2	100Ω	brown black brown brown	brown black black black brown
❑	1	10Ω	brown black black brown	brown black black gold brown

TempMaster Connection Options

These diagrams show three different ways that the *TempMaster Mk3* can be connected up to control the temperature of a fridge, freezer or heater set-up.

Which one you use will depend on whether your fridge/freezer/heater operates from 230VAC or 12V DC, and also whether you will be running it from the AC mains or from a battery supply.

Option A shows the simplest arrangement, where a 230VAC fridge/freezer or heater is to be operated directly from the mains supply.

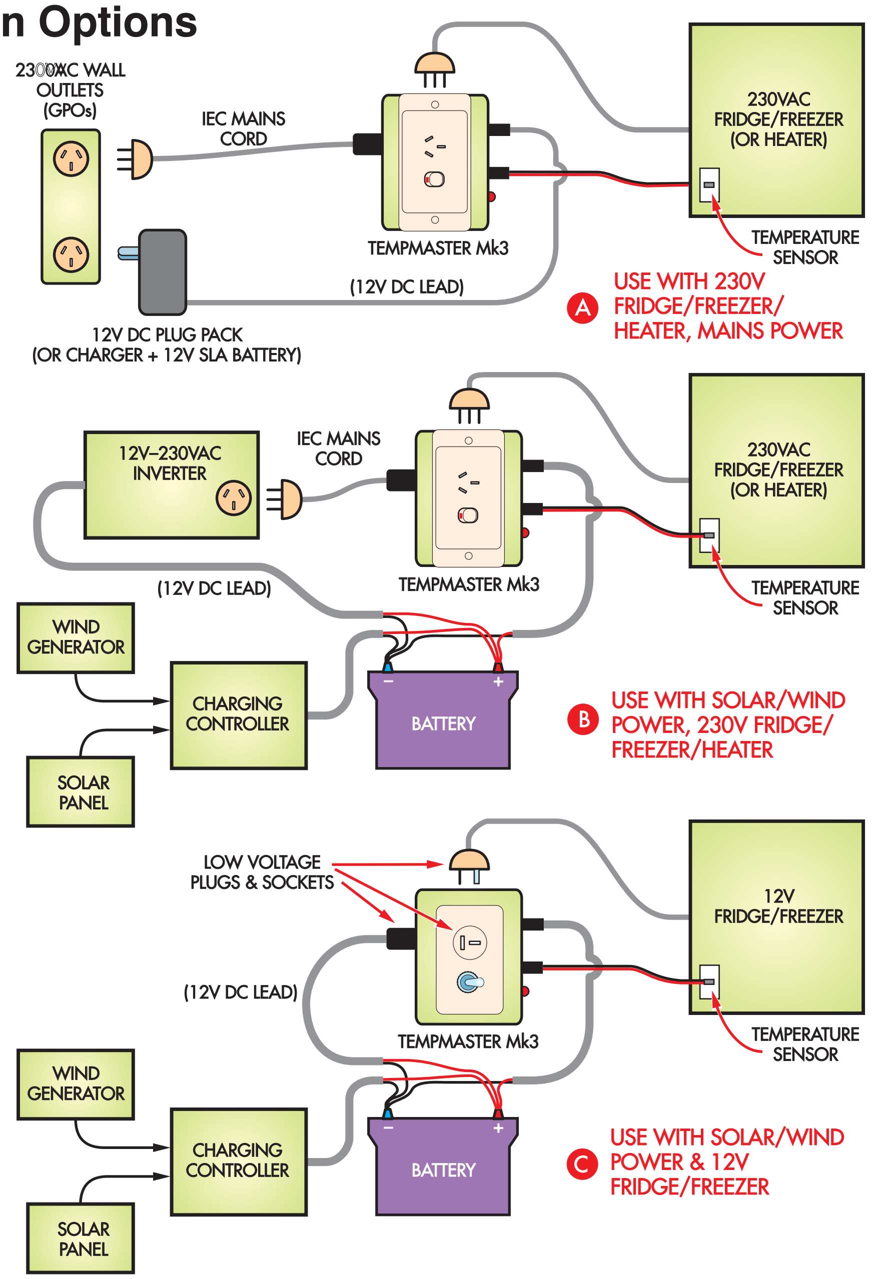
The 12V DC needed by the *TempMaster* itself can be supplied either by a small 'plug pack' DC supply or from a 12V SLA battery, which is kept 'topped up' by a suitable charger.

Option B shows how a 230VAC fridge/freezer or heater can be connected to a 12V/230VAC power inverter, in a home or building which relies on solar or wind generated power.

The *TempMaster* itself can be powered from the main battery, along with the power inverter used to operate the fridge/freezer/heater.

Because there is no current whatever drawn from the *TempMaster's* IEC mains input socket when the *TempMaster* has switched off the power to the fridge/freezer/heater, the inverter should be able to drop back to 'sleep' mode at these times.

Option C shows how to connect things up when the *TempMaster* is to be used with a 12V fridge/freezer and a solar power system. In this case, you MUST replace both of the *TempMaster's* 'mains' connectors with suitable low voltage plugs and sockets, to make sure that they can't be accidentally connected to 230VAC.



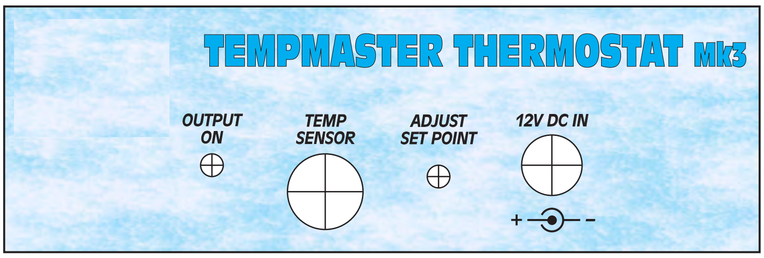
If you mount the thermostat case on the wall just behind the fridge/freezer or heater, the plug on the end of the ribbon cable can be plugged into CON2 on the lower front of the case to complete the job.

Now you can unplug the power cord of the fridge/freezer/heating cabinet from its original GPO socket and plug it instead into the GPO on the top of the *TempMaster Mk3*. When you connect the *TempMaster Mk3's* own IEC mains connector to the original GPO via a suitable IEC mains cable, the complete system will begin working. (You do have to flick the switch on the *TempMaster Mk3's* GPO to the 'on' position, of course!)

If you want to make sure that the thermostat is holding the fridge/freezer/heater to the temperature you want, this can be done quite easily using a thermometer placed inside the cabinet for a while.

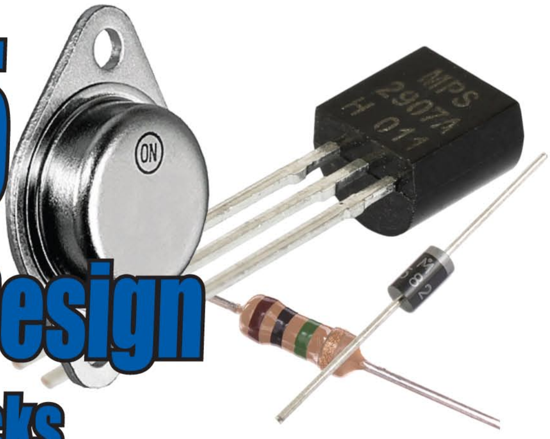
You can see when the *TempMaster Mk3* is switching power to the compressor or heater simply by watching LED1.

If you need to adjust the average temperature up or down, this is done quite easily by adjusting trimpot VR1 using a small screwdriver, through the small hole in the front of the case (between the holes for CON1 and CON2).



Full-size artwork for the *TempMaster Mk3* front panel, which mounts on the box side. The GPO fastens through the top of the box.

Teach-In 2015



Discrete Linear Circuit Design

Part 7: Heat and more building blocks

by Mike and Richard Tooley

Welcome to *Teach-In 2015*. This series is aimed at anyone wishing to develop a detailed understanding of linear discrete semiconductor devices and how they are used in a diverse range of circuits. We hope you will join us on this exciting voyage of discovery! Each part of our *Teach-In 2015* series is devoted to a different aspect of discrete linear circuit design, such as modelling and simulation, measurement and testing, noise and distortion. In last month's instalment, *Knowledge Base* introduced you to two useful circuit building blocks in the form of constant current and constant voltage sources. *Discover* was devoted to power and power measurement while our regular practical feature, *Get Real*, described the tests and measurements that we made on our simple tone control and revealed that we met or exceeded most, if not quite all, of our original design objectives.

Introduction

In this month's *Teach-In 2015*, *Get Real* describes the construction and use of a simple VU-meter. *Discover* examines heatsink design, and *Knowledge Base* introduces three useful circuit building blocks: the current mirror, the differential amplifier and the V_{BE} multiplier.

Discover: Heat and heat dissipation

The electronic designer is often faced with the task of designing circuits that will operate safely and reliably over a wide range of ambient temperature. Heat is the enemy of most electronic circuits, and getting rid of it can be an important consideration when dealing with semiconductors and other devices that might need to operate at appreciable levels of power.

Determining temperature rise

With increasing levels of power output, consideration needs to be given to the heat generated by a semiconductor device as well as ways of effectively dissipating it. The primary aim is ensuring that the temperature experienced by the semiconductors (particularly those used in the output stage) never becomes excessive. Here we need to consider the heatsink as well as the mounting arrangement used to attach a device to it.

The ability of a heatsink and associated mounting arrangement to dissipate heat can be measured in terms of the increase in semiconductor junction temperature above the ambient (surrounding air) temperature. A perfect heatsink and mounting arrangement would keep the junction temperature the same as that of the surrounding air. In practice, a rise in junction temperature is unavoidable. Paramount in all of this is the need to ensure that the manufacturer's maximum junction temperature, T_{Jmax}, is never exceeded when the device is dissipating its maximum rated power. To assist designers, manufacturers provide a de-rating characteristic that can be used to determine the maximum permissible power dissipation when the junction temperature exceeds its specified value for normal operation (see Fig.7.1).

Fig.7.1 shows that the device in question (a common TIP41 NPN silicon power transistor) is fully rated for the total (collector plus base) power dissipation of 65W for case temperatures of up to 25°C but, above this temperature, the device must be de-rated in a linear fashion until the maximum junction temperature of 150°C is reached (at which point the total power dissipated should be zero!). The characteristic is useful in various ways. For example, if the case temperature is allowed to run up to 75°C the total power dissipation should not

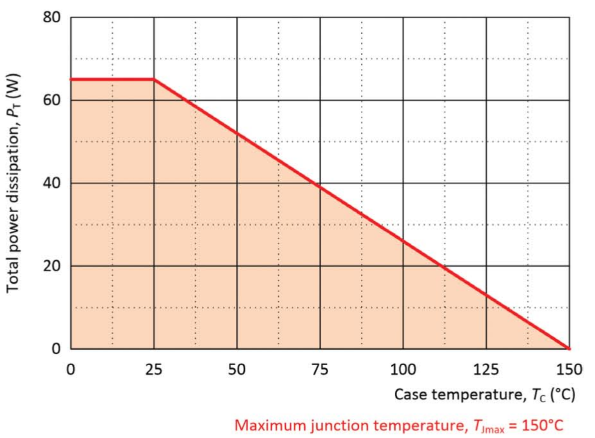


Fig.7.1 Thermal de-rating characteristic for a TIP41 medium power transistor

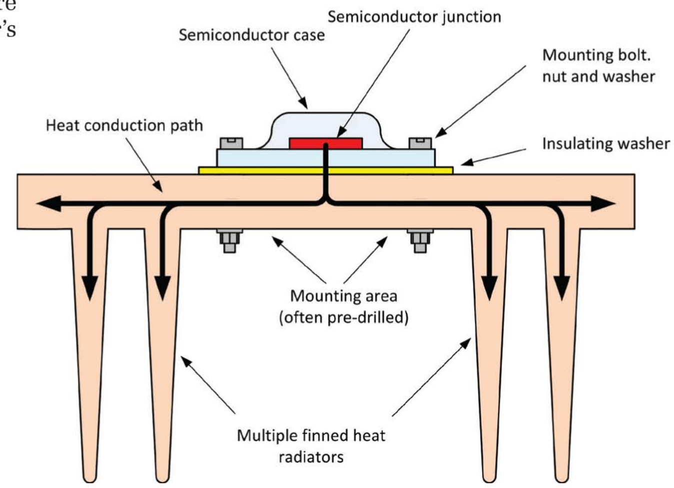


Fig.7.2 Typical heatsink and mounting arrangement for a TO3 packaged semiconductor

be allowed to exceed 40W. Conversely, if the device is expected to dissipate 50W its case temperature should not be allowed to rise to more than about 50°C.

Thermal resistance

An object's ability to resist heat flow is referred to as its 'thermal resistance'. Thermal resistance is the opposite of thermal conductance; a good heat conductor would exhibit a very low thermal resistance, while a poor heat conductor would exhibit a very high value of thermal resistance. Clearly, what we need for an effective heat dissipater is a very low value of thermal resistance between the semiconductor junction (encapsulated inside the semiconductor's package) and the air that surrounds it. We can achieve this with an appropriate selection of semiconductor package, heatsink and mounting hardware.

Calculating temperature rise

When determining mounting and heatsinking arrangements, one of the first questions that we might need an answer to is what temperature rise to expect above that of the surrounding environment. The temperature rise, ΔT, *above ambient* will be given by:

$$\Delta T = \theta_T \times P_T$$

where P_T is the total power dissipated by the semiconductor device(s) and θ_T is the *total* thermal resistance of the heatsink and mounting arrangement.

To put this into context, assume that we have a transistor that is dissipating a power of 6W and that the total thermal resistance present is 11.5°C/W. The temperature rise *above ambient* would amount to (6 × 11.5)°C or 69°C. If the ambient temperature had been 25°C this would result in a junction temperature of (69 + 25)°C or 94°C which should, perhaps, begin to sound a few alarm bells!

Determining the junction temperature

The temperature of the semiconductor junction can be determined from the following relationship, where T_A is the ambient temperature:

$$T_J = (P_T \times \theta_T) + T_A$$

Table 7.1 Thermal resistances for various transistor case styles

Case style	θ_{JC} (°C/W)	θ_{JA} (°C/W) (See note)
TO92	40 to 60	200 to 350
TO126	3 to 10	83 to 100
TO220	1.5 to 4	60 to 70
TO202	6 to 13	62 to 75
TO218	1 to 1.6	30 to 45

Note: Unmounted semiconductor package and no heatsink present

Typical maximum junction temperatures quoted by manufacturers range from about 150°C to 200°C depending upon application and package style. Beyond the quoted maximum junction temperature there is a risk of permanent damage to the semiconductor device in question.

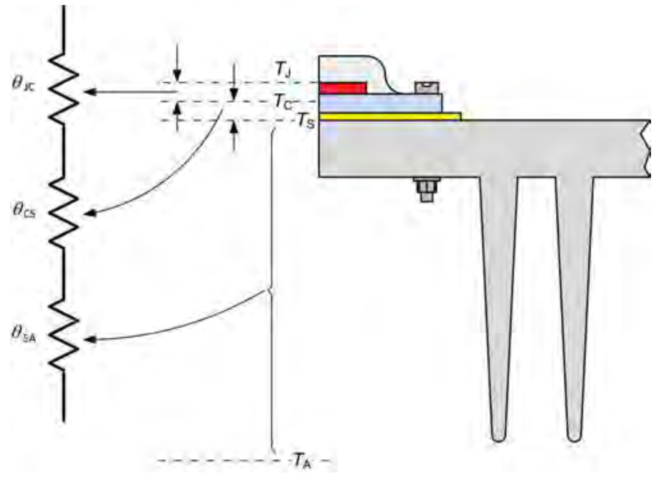


Fig.7.3 The three thermal resistances present in Fig.7.2

Determining thermal resistance

Earlier, we said that θ_T was the total thermal resistance present. This thermal resistance arises from several sources, as depicted in Fig.7.2, which shows a typical TO3-case-style transistor mounted on a finned heatsink. Heat is conducted away from the semiconductor junction to the outer case of the TO3 package and then, via an insulating washer to the surface of the heatsink. From this point, heat is conducted to the extremities of the fins, where it is radiated into the surrounding air space. Thus, the total thermal resistance, θ_T, present in Fig.7.2 is the sum of three individual thermal resistances:

1. The thermal resistance that exists between the semiconductor junction and the case of the transistor (ie, the thermal resistance *inside* the transistor package), θ_{JC}
2. The thermal resistance of the insulating washer (ie, the thermal resistance from case to surface), θ_{CS}
3. The thermal resistance between the surface of the heat radiator and the space surrounding it (ie, the thermal resistance between the surface and ambient), θ_{SA}

Fig.7.3 shows these three thermal resistances together with the temperatures that exist at each point in the arrangement shown in Fig.7.2. It should be apparent that the three thermal resistances shown in Fig.7.2 actually appear 'in series' and we can use a simple electrical analogy to represent the thermal 'circuit' in electrical terms, as shown in Fig.7.4. From this arrangement we can conclude that the total thermal resistance, θ_T, is given by:

$$\theta_T = \theta_{JC} + \theta_{CS} + \theta_{SA}$$

Notice that the total thermal resistance present is actually the same as the thermal resistance from junction to ambient, θ_{JA}. Thus:

$$\theta_{JA} = \theta_{JC} + \theta_{CS} + \theta_{SA}$$

The complete electrical equivalent circuit of the heatsink arrangement is shown in Fig.7.4. Note that the source of power (P_{TOT}) is the semiconductor device and the 'potentials' at the two extreme ends of the series chain of thermal resistances are T_J (junction temperature) and T_A (ambient or surrounding air temperature). Table 7.1 shows some typical thermal resistances for various transistor case styles, which show how an unmounted device performs when no heatsink is present.

Worst-case conditions

In the design of electronic equipment it is prudent to plan for the worst-case conditions, ensuring that the absolute maximum junction temperature, T_{Jmax}, is not exceeded when the total power dissipation and ambient temperature jointly reach their maximum working values. As an example, consider the following scenario:

A transistor has an absolute maximum junction temperature rating of 150°C and a thermal resistance from junction to case of 1.0°C/W. If the device is fitted with a washer and mounting kit having a thermal resistance of 1.25°C/W and a heatsink of 2.75°C/W, determine whether the maximum ratings are exceeded when the total power dissipation reaches a maximum of 25W at an ambient temperature of 40°C. Applying the equations that we met earlier gives:

$$\theta_{JA} = \theta_{JC} + \theta_{CS} + \theta_{SA}$$
$$= (1.0 + 1.25 + 2.75) = 5°\text{C/W}$$

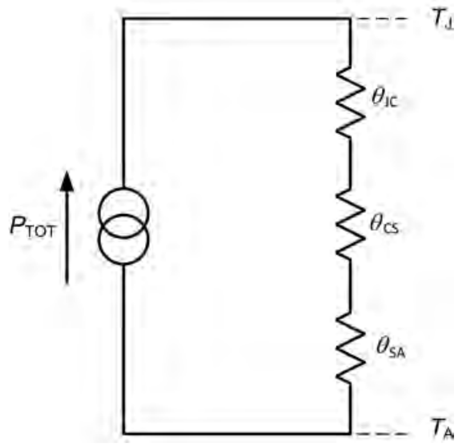
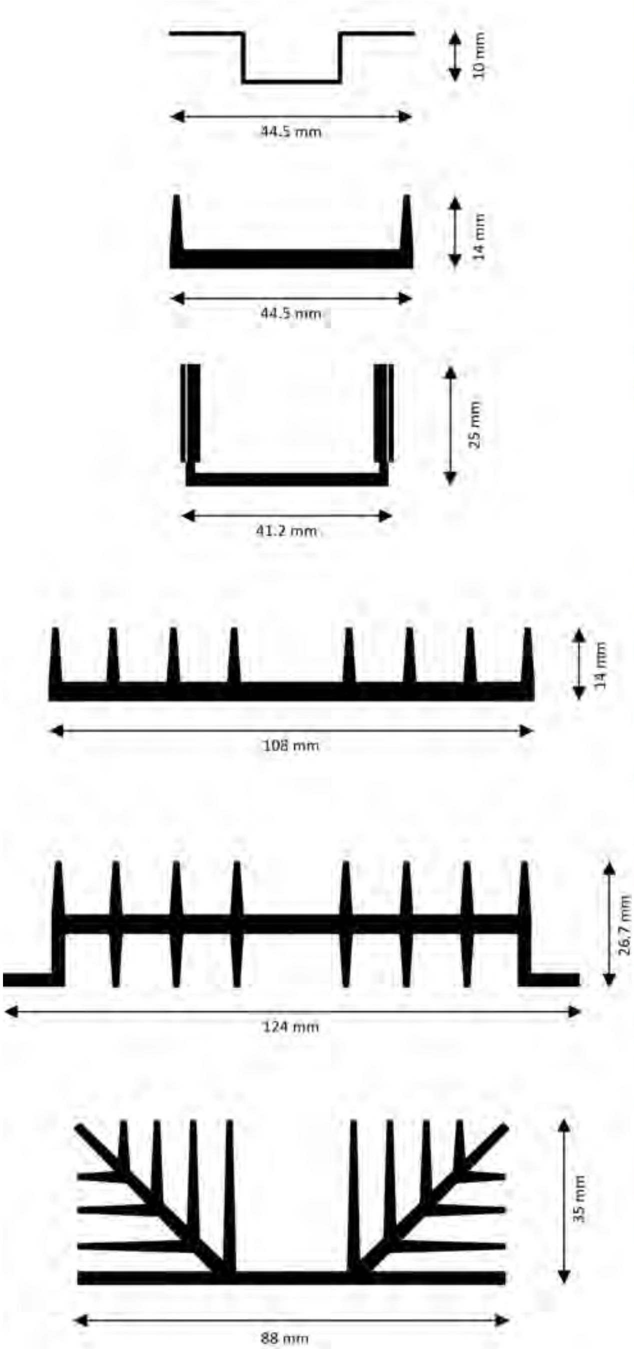


Fig.7.4 Equivalent circuit showing the three thermal resistances in Fig.7.3



Length (mm)	Thermal resistance (°C/W)
17.8	20.0

Length (mm)	Thermal resistance (°C/W)
32	14.0

Length (mm)	Thermal resistance (°C/W)
38	7.1

Length (mm)	Thermal resistance (°C/W)
50	5.0
75	3.9
87.5	3.5
100	3.2
150	2.6

Length (mm)	Thermal resistance (°C/W)
75	2.72
100	2.3
150	1.85

Length (mm)	Thermal resistance (°C/W)
75	1.65
100	1.40
125	1.20

Fig.7.5 Some typical heatsink cross-sections and thermal resistances

If the transistor is dissipating 25W, the junction temperature will rise to:

$T_J = (P_T \times \theta_T) + T_A$
$= (25 \times 5) + 40 = 165°C$

This exceeds the 150°C absolute maximum junction temperature rating by 10%, and so the designer should either reduce the power dissipation to a safe value or reduce the thermal resistance of the heatsink arrangement (or both).

Determining heatsink specifications

The designer often has to determine the required heatsink specifications given the absolute maximum junction temperature, thermal resistance from junction to case, maximum expected ambient temperature etc. To do this, we need to rearrange the equation to make θ_{SA} the subject of the equation. Thus:
$\theta_{SA} = \theta_{JA} - (\theta_{JC} + \theta_{CS})$

The value obtained for θ_{SA} will be the minimum acceptable rating for the required heatsink and, in practice, we would choose a component with a higher rating to allow for a margin of safety. From the above, we know that:

$\theta_{JA} = (T_J - T_A) / P_T$

Thus: $\theta_{SA} = ((T_J - T_A) / P_T) - (\theta_{JC} + \theta_{CS})$

To put this into context, let's assume that we need to determine the minimum acceptable thermal resistance rating for a heatsink that will be used with the transistor that we met earlier:

$\theta_{SA} = ((150 - 40) / 25) - (1 + 1.25)$
$= (110 / 25) - 2.25 = 2.15°C/W$

In practice, a substantial heatsink of around 1.9°C/W would be suitable for use in this application.

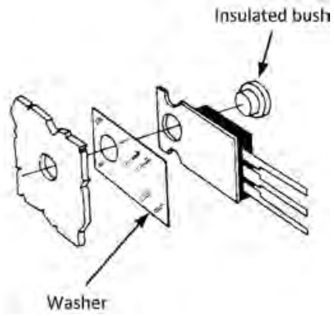


Fig.7.6 Mounting arrangement for a TO220 semiconductor package

Practical heatsink arrangements

A selection of commonly available heatsink cross sections is shown in Fig.7.5. These range from a simple folded U-section metal plate with a thermal resistance of 20°C/W to a complex aluminium alloy extrusion with a thermal resistance of 1.2°C/W. Lower values of thermal resistance can be obtained with the use of forced-air cooling using a fan. For example, forced-air cooling will typically reduce quoted thermal resistances by as much as 50% when an air flow of around 200 litres per minute is present.

Where forced-air cooling is not required, natural convection airflow can be enhanced by the proper placement of heatsinks and other heat producing components. Since warm air rises, vertical surfaces tend to transmit heat to the air better than comparable horizontal surfaces. The hottest devices should be located on the upper side of a horizontally mounted PCB or close to the upper edge of a vertically mounted PCB.

A typical mounting arrangement for a TO220 semiconductor package is shown in Fig.7.6. In many cases, the tab of the device is connected to one of three terminals (often the collector or drain) in which case a mica or thermally conductive plastic washer must be fitted. Note also that an insulated bush must be used in order to prevent the mounting bolt shorting the metal tab to the heatsink.

It's also important to remember that the thermal resistance of the mounting kit used with a semiconductor device can have a major effect on the efficiency of the heat conduction from the surface of the case to the heat radiator. Special thermally impregnated washers have significantly lower thermal resistance than simple mica washers. Thermally conductive silicone grease should NOT be used with this type of washer.

Knowledge Base: More building blocks

This month, we will be looking at some more useful circuit building blocks in the form of the current mirror, differential amplifier and V_{BE} multiplier. As with the previous circuit building blocks

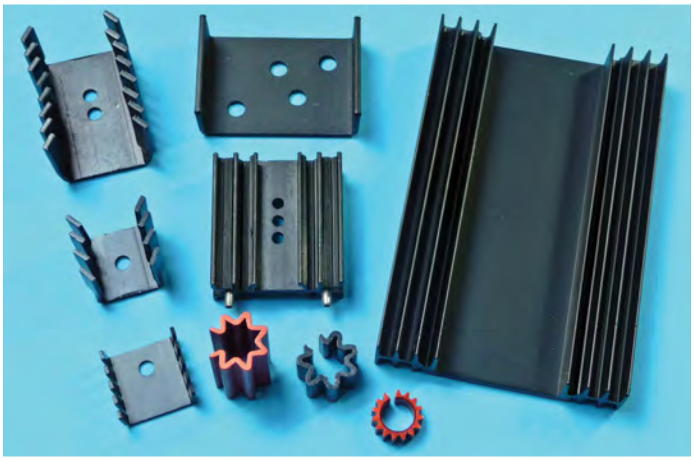


Fig.7.7 A selection of common heatsinks with thermal resistance ranging from 4°C/W to 63°C/W

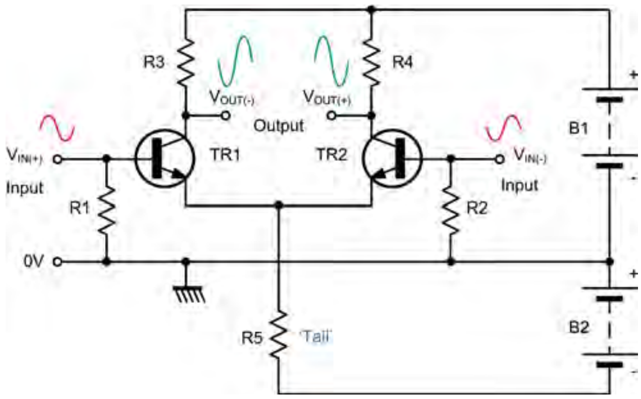


Fig.7.9 A conventional differential amplifier

described in this series, we've provided a set of models that can be used with the Tina Design Suite. These can be downloaded from the *EPE* website and will enable you to test, modify and experiment with each of the circuit arrangements that we've described here.

The current mirror

As its name suggests, the current mirror provides a means of closely matching two currents. This is useful in circuits where two devices need to be supplied with the same current. Furthermore, when the current increases in one branch the other branch will experience a similar increase, and vice versa.

The circuit of a current mirror is shown in Fig.7.8. The input stage formed by R1 and TR1 converts the input current (IC1) into a voltage (V_{CE1}) and an output stage formed by TR2 that converts an input voltage (V_{BE2}) into a current (IC2). Note that $V_{CE1} = V_{BE1} = V_{BE2}$ and so, with identical devices for TR1 and TR2, the circuit effectively replicates the current supplied to the first active device (TR1) in the second active device (TR2).

A particularly useful feature of the current mirror is a relatively high output resistance, which helps maintain the output current constant, regardless of load conditions. Another characteristic of the current mirror is a relatively low input resistance. This helps keep the input current constant regardless of drive conditions. Note that the replicated current is, in many cases, a signal current superimposed on a static (or quiescent) current. The current mirror is often used to provide bias currents and active loads in small and large signal amplifier stages.

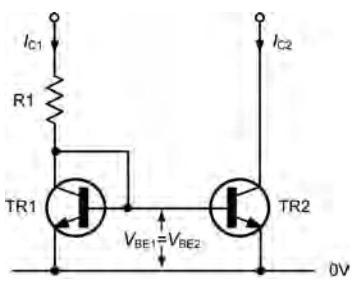


Fig.7.8 A current mirror

The differential amplifier

A conventional differential amplifier has two inputs and two outputs (as shown in Fig.7.9) and neither of its inputs is grounded. The differential amplifier shown in Fig.7.9 consists of two identical common-emitter stages with their emitters linked and a common 'tail' resistor (R5). As its name suggests, the output of the differential amplifier is proportional to the difference in voltage at its two inputs. The circuit shown in Fig.7.9 can be more closely balanced (compensating for any small difference in transistor current gain) by including a pre-set resistor in the 'tail', as shown in Fig.7.10. When a signal is applied to the two inputs the circuit effectively behaves like a 'see-saw', a positive-going input voltage applied to the base of TR1 results in a negative-going voltage at its collector, while at the same time, the corresponding negative-going input voltage at the base of TR2 produces a positive-going voltage at its collector.

A further improvement to the circuit shown in Fig.7.11 is the inclusion of a constant current source (see last month's *Teach In 2015*) in the 'tail' circuit. This constant current source (formed by TR3 and its associated components) supplies the combined emitter current of the two devices. The constant current source shown in Fig.7.11 effectively holds the total emitter current constant so that, when the emitter current in one device increases the emitter current in the other device decreases by an equivalent amount, and vice versa.

The differential arrangement that we've just described provides us with a useful means of comparing the output signal of an amplifier with its input, as shown in Fig.7.12. This 'comparator' arrangement helps to stabilise the overall gain of an amplifier as well as improving its linearity. We will be looking at this topic in further detail in Part 8 of our *Teach-In 2015* series.

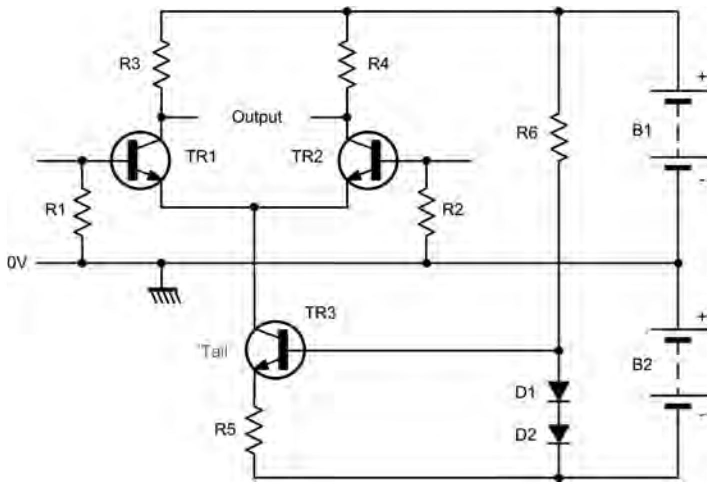


Fig.7.11 An improved differential amplifier with a constant current 'tail'

Differential amplifier as a phase splitter

The differential amplifier also provides us with a neat way of splitting a signal into two signals having opposite phases (one in phase with the input signal and one that is 180° out of phase with the input signal). In such a case, one of the two inputs of the amplifier is effectively grounded (via C2) and the ground-

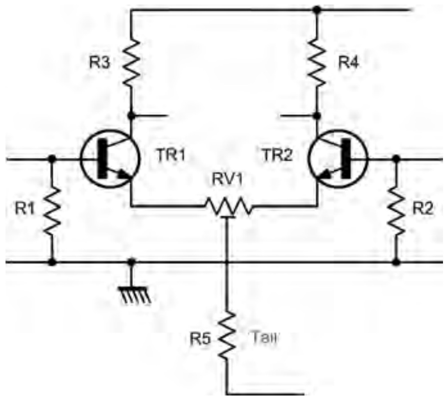


Fig.7.10 A differential amplifier that can be accurately balanced

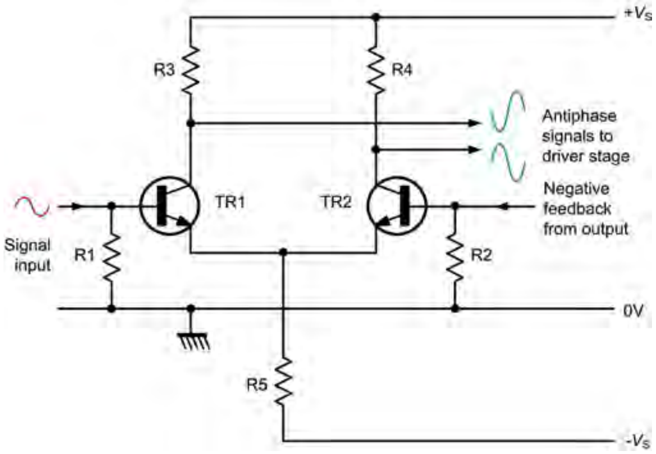


Fig.7.12 Using a differential amplifier as a comparator

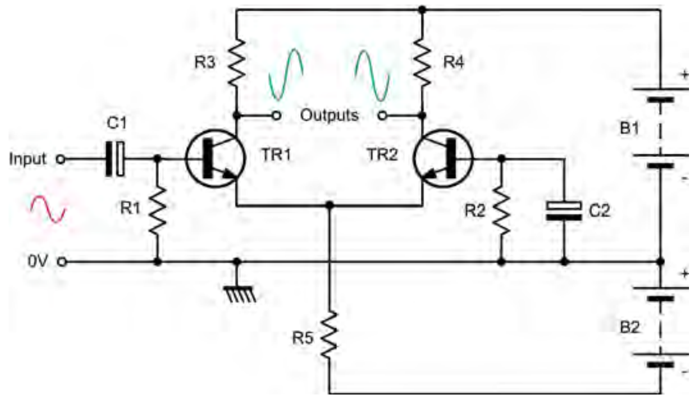


Fig.7.13 A differential amplifier used as a phase splitter

referenced signal is applied to the other input. The two anti-phase outputs are then taken from the two collectors (as before). Fig.7.13 shows this arrangement (note that our *Get Real* project this month makes use of an alternative form of phase splitter using just one transistor).

The V_{BE} multiplier

Our final circuit building block this month (see Fig.7.14) provides us with a handy and very effective means of stabilising the bias voltage applied to a pair of transistors in the output stage of an amplifier. When conducting, the voltage at the base-emitter junction of a silicon transistor (V_{BE}) is assumed to be around 0.7V. By deriving this voltage from the potential divider formed by R1 and R2 in Fig.7.14, the collector voltage will be held at a constant voltage, V_{CE}, given by:

$$V_{CE} = \left(\frac{R1 + R2}{R1}\right) \times V_{BE} = 0.7 \times \left(\frac{R1 + R2}{R1}\right)$$

Thus, to produce a stabilised bias voltage of, say, 1.27V, and with R1 of 1kΩ we would need to calculate the value of R2 using:

$$R2 = \left(\frac{R1 \times V_{CE}}{0.7}\right) - R1 = \frac{1.27}{0.7} - 1 = 0.814\text{k}\Omega$$

Thus, a value of around 820Ω would be satisfactory. In practice, we would usually require the bias voltage to be adjustable and so the two fixed resistors would be replaced by a pre-set potentiometer of around 2kΩ, as shown in Fig.7.15.

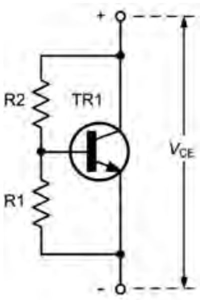


Fig.7.14 A V_{BE} multiplier

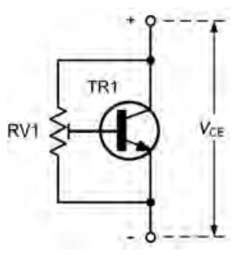


Fig.7.15 Making the V_{BE} multiplier adjustable

Get Real: A simple VU-meter

Our fourth *Get Real* project is a simple VU-meter. This module will provide you with a reliable means of measuring the signal level (in 'volume units') present in a system. VU-meters indicate signals relative to a standard defined as '0VU', equivalent to signal of 1.228V_{RMS} into a 600Ω load, and equivalent to a power level of about 2.5mW (+4dBm). The range indicated by a VU-meter is designed to show a range of signal levels that can be standardised for recording and broadcast purposes; '–3dB' usually appearing centre-scale with the lowest and highest marked indications of '–20dB' and between '+3dB' and '+5dB', respectively (see Fig.7.16).

It is important to be aware that a VU-meter is designed to indicate the *average* level of a signal without responding to the sudden peaks and troughs that are often present in speech and music. In applications such as broadcasting, where a particular level of amplitude must not be exceeded, it is usually more important to indicate the peak rather than the average level of a signal. This can be achieved using an alternative instrument known as a 'peak programme meter' (PPM). Calibration and specification of a VU-meter is usually carried out using a sinewave at 1kHz and its frequency response will usually extend from around 20Hz to 20kHz within ±1dB, and with a step-response time of around 300ms.

Circuit description

The complete circuit of our simple VU-meter is shown in Fig.7.17. The circuit uses a single transistor, TR1, which acts as a phase-splitter producing signals of equal amplitude but of opposite phase at its collector and emitter. These two anti-phase signals are applied to rectifiers D1 and D2. D1 acts on negative-going half-cycles, while D2 acts on positive-going half cycles of the input waveform applied to the base of TR1. This arrangement ensures that the VU-meter responds to both negative- and positive-going half cycles of any applied waveforms that might be asymmetric in nature (eg, speech or music). Such an arrangement would not be needed if the circuit was only to be used with symmetrical signals (such as pure sinewaves).

In order to improve linearity at low signal levels a small amount of forward bias is applied to the two rectifier diodes by means of an adjustable pre-set resistor, RV2. The voltage at the slider of RV2 is adjusted to the point at which D1 and D2 just start to conduct (around 0.6V). The forward-biased silicon diode, D3, ensures that this voltage never exceeds 0.7V, or so. The DC output from D1 and D2 is summed via R5 and R6 and averaged by means of C5. The combined DC output current is fed to a moving coil meter via an adjustable pre-set resistor, RV3.

Construction

Our prototype printed circuit board (PCB) was designed to be built into a small separate enclosure or incorporated into a larger enclosure along with other circuitry; it measures just 122mm × 30.5mm. As with our other projects, the PCB component layout (Fig.7.18) and copper track layout (Fig.7.19) were produced using Circuit Wizard. The board can be purchased, ready drilled, from *EPE PCB Service*, code 908.

Components

General

1 PCB, code 908 available from the *EPE PCB Service*, size 122mm × 30.5mm
3 PCB mounting 2-way terminal blocks
1 PP3 battery connector
1 SPST on/off switch

Fixed resistors (all are 0.25W 5%)

1 680Ω (R1)
2 4.7kΩ (R2 and R7)
2 470Ω (R3 and R4)
2 2.2kΩ (R5 and R6)

Fig.7.16 A typical VU-meter scale

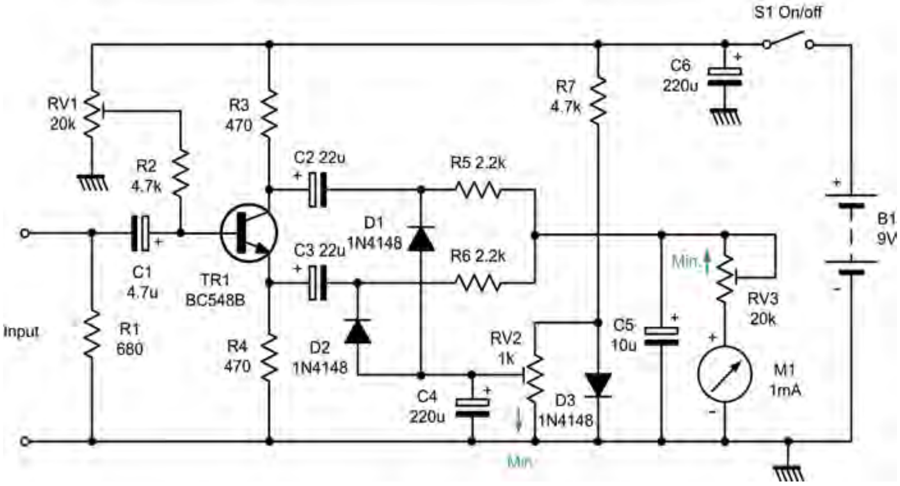


Fig.7.17 The complete circuit of the simple VU-meter

Pre-set resistors
2 20kΩ (RV1 and RV3)
1 1kΩ (RV2)

Capacitors
1 4.7µF (C1)
2 22µF (C2 and C3)
2 220µF (C4 and C6)
1 10µF (C5)
All capacitors are 16V or 25V radial types

Semiconductors
1 BC548B (TR1)
3 1N4148 (D1, D2 and D3)

Meter
1mA meter movement with VU-scale

Note that if you are building two VU-meters for stereo operation you will need two sets of components.

Calibration

Set-up and calibration is quite simple using only a sinewave signal generator and a multimeter or oscilloscope. With no signal applied, RV1 is first adjusted so that the voltage at the collector of TR1 is approximately 6V. The voltage at the emitter of TR1 should then be measured and this should be approximately 3V. Next RV2 and RV3 should both be set to minimum (see Fig.7.17). With no signal applied, RV2 should slowly be advanced until the pointer of the meter just starts to lift from zero. RV3 should then be set to mid-position.

A multimeter or oscilloscope should then be connected to read the input voltage and a sinewave signal at 1kHz should be applied. The output level from the signal generator should be set to 1.3V_{RMS} (using a multimeter) or 3.7$V_{pk\text{-}pk}$ (using an oscilloscope). RV3 should then be adjusted so that the meter indicates a reading of exactly '0VU'. Finally, when using the simple VU-meter it is important to remember that it has an input impedance of 600Ω and is designed for use in a system with 600Ω impedance, or lower. If necessary, the input impedance can be raised to around 5kΩ by removing R1.

Next month

In next month's *Teach-In 2015*, *Discover* will be devoted to high power amplifiers, and the Darlington and Sziklai pair configurations that are commonly found in them. *Knowledge Base* will introduce negative feedback and explain how it provides a useful and very effective way of making an amplifier stable and predictable.

Errata

Thank you to Dave Reeves who spotted an error on page 42 of June's *Teach-In 2015*. In the third column, under the heading 'Signal-to-noise ratio', the left-hand side of the first equation reads '$(S+N)/N=$', it should be simply: '$S/N=$'.

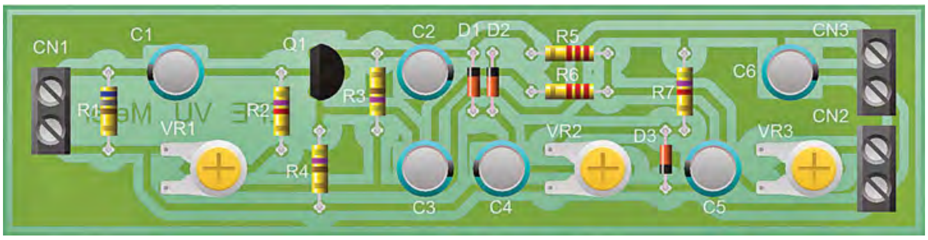


Fig.7.18 PCB component layout shown using Circuit Wizard's 'real world' view

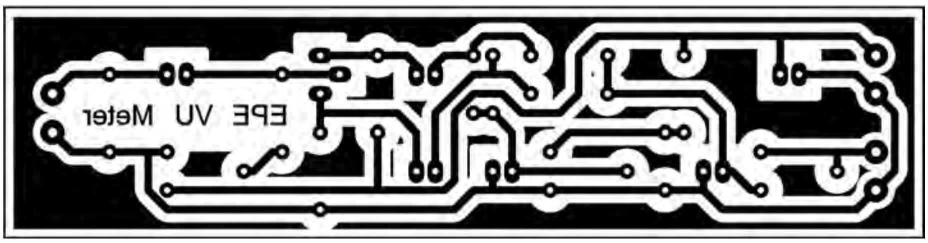


Fig.7.19 PCB track layout

NET WORK

by Alan Winstanley

An echo of the future

As a teenager of the 1970s, your scribe earned his pocket money in a WHSmith newsagency as a Saturday lad, where many magazines – including *Everyday Electronics* and *Practical Electronics* – jostled for space on the stands alongside newfangled Pentel rollerball pens, vinyl records and Dymo tape labellers. A slab-sized demonstrator Rockwell LED calculator was chained to the countertop. (Sentimentalists yearning for that era might enjoy the *EPE Chat Zone* thread and video at: **http://www.chatzones.co.uk/discus/messages/7845/15089.html.**)

Plenty of hobby electronics magazines on sale but no barcodes or IT in this 1980s newsagent (from the Rising Damp movie – © British Lion Films)

All merchandise was priced individually, which involved the writer earnestly printing out endless coils of sticky price labels (hopefully bearing the correct price) and applying them by hand to each and every product. Point-of-sale analysis meant punching the price into the correct category on a cash register, which the store manager totalled up expectantly later in the day. Credit card transactions involved signing slips of paper with no other security measures evident at all, depositing the payslips into a bank and crossing one's fingers afterwards.

As barcodes gradually appeared, shoppers initially hated the idea of seeing prices on shelf-edge labels instead of little price stickers. Items could now be tagged uniquely and scanning systems could check stocks and prices on a central database. The concept of Stock Counting Units (SKUs) to manage inventories caught hold, and refinements in technology saw barcodes appearing on everything from Polo mints to palletloads of potatoes. Gradually, IT systems and Electronic Data Interchange (EDI) enabled stock control, supplier ordering and logistics to be integrated more seamlessly. E-commerce gradually followed, which made the purchasing process even more accessible until electronic transactions became routine.

Amazon's magic wand

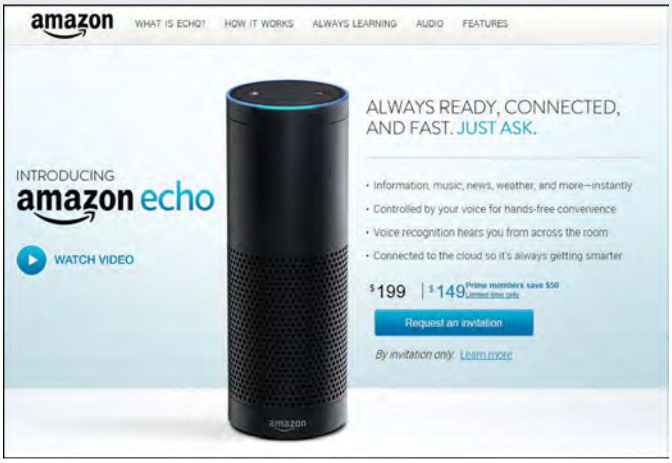
Despite all this clever technology, commerce still revolves around customers somehow placing orders and suppliers fulfilling them. As a sign of future trends, Amazon's Dash wand (see *Net Work*, July 2014) promised last year to scan product barcodes in the home and re-order them through a buyer's AmazonFresh account with just a button-click or two. Interestingly, the gadget also incorporated voice recognition. The Dash wand has been a low-key development shipped by invitation only and it's described in more detail at: **https://fresh.amazon.com/dash**. If nothing else, the Dash wand gave us an insight into the way Amazon was looking into the future of home shopping.

As mentioned last month, Amazon now hopes to overcome another hurdle in the buying process with Amazon Dash buttons, which are becoming available to their Prime customers in the US. These small stick-on fobs are dedicated to a single product and a tap of the button emits a Wi-Fi signal to replenish supplies; Amazon duly fulfils the order and delivers products to the door. The Dash idea may also find its way into appliances such as washing machines, so you need never run out of soap or conditioner again. Of course, the button-press is a painless experience for buyers, since no hard cash changes hands and prices are merely data fields shown in a shopping cart.

Ian Crouch, blogging for **NewYorker.com**, likened Amazon Dash buttons to a Skinner box, the laboratory apparatus that conditions rats into pressing a certain lever in order to be rewarded with food. Whether the Dash push-button proves viable remains to be seen, but it is another example of how shopping habits are being shaped for the future by embracing Wi-Fi devices and the Internet. Perhaps Dash buttons will be most useful for topping up trivial nuisance consumables where buyers don't feel the need to shop around much, but buyers might also 'do a Tesco' and eventually rebel against the idea of being locked into pricey suppliers; for my part, I recently tried ordering a 1,000-pack of tea bags from Amazon UK, but soon cancelled it as the postage charges made the idea non-viable: I could buy much cheaper locally. At least Amazon Dash buttons will only place one order at a time, so multiple clicks of the button won't result in truckloads of toilet rolls arriving next day. A Youtube video at: **https://www.youtube.com/watch?v=NMacTuHPWFI** gives the low-down on Amazon Dash and how Amazon sees its push-button product fitting into our daily lives.

You can call me Alexa

Amazon's foray into the kind of voice recognition and AI seen in last year's Dash Wand is being taken a step further with Amazon Echo, an example of home hardware featuring Amazon's new voice assistant called Alexa, a voluble lass who is Amazon's answer to Apple's Siri, Microsoft's Cortana and Google Now. Echo is a black cylindrical device with built-in speaker that hooks via Wi-Fi to an Amazon account. Its voice recognition enables users to add items to a shopping list or a To-Do List, or they can generally interact with Alexa in the same way that rival AI voice apps obey commands or answer a multitude of questions. Echo



Amazon Echo features its voice assistant Alexa; it connects wirelessly to the Amazon cloud

Samsung offers its Artik environment for controlling IoT devices using low-power miniature modules

also has a remote control to allow operation from across the room, and it also offers apps for Fire OS, Android and the desktop computer. It claims to be Belkin Wemo and Philips Hue compatible (see *Net Work*, May 2014), allowing remote control of those networkable devices, and it's an alarm clock, news and weather forecast presenter as well. Furthermore, network support for *If This Then That* is now possible; IFTTT is a cloud-based macro system offering a dedicated Alexa 'channel' to trigger a desired 'action' (*Net Work* April 2014). As Amazon also sells video and music, entertainment can be lined up via Echo and streamed to your smart TV or tablet. Amazon Echo is currently available by invitation only in the US, but the serious capabilities of this fascinating piece of hardware signpost the way ahead and Siri, Cortana and Google Now may have to make way for a new voice among their numbers. More details are at: **www.amazon.com/echo**

Break out the Brillo

It is sometimes easy to forget how fast Internet technology is moving along: almost 20 years ago owners of the new Nokia 6110 mobile phone could personalise their phone 'online' using their phone's WAP-compliant web browser to view a clunky text-based WAP page over a very slow network. They could then waste half an hour ordering (of all things) a bespoke Nokia microphone bezel overprinted with their initials. Such phones also lent themselves to controlling devices via the phone network – for example, in the March to May 2007 issues of *EPE* we published an SMS controller that utilised text messaging in a simple control system. Google is now toying with the idea of producing a customised, modular smartphone fit for the 21st Century: its Project Ara (**http://www.projectara.com**) adopts a Lego-brick approach to building an ideal phone, where users can pick and mix sensors, cameras and shells to make a fully bespoke mobile phone that allows owners to better 'express themselves', as their PR blurb put it. It draws on the Phonebloks concept (**https://phonebloks.com/en**) and Google is also excited about the prospect of 3D-printing some of the parts. Project Ara lives in Google's Advanced Technology and Projects (ATAP) group and more details can be found on Youtube at: **https://www.youtube.com/watch?v=zG_uwDqLsZY**.

Google has also announced an operating system optimised for IoT devices. It is dubbed Project Brillo and is a form of 'Android Lite' dedicated to future IoT devices running in the smart home. A cloud-related protocol called Weave would allow Brillo-compatible devices (eg, Nest Protect) to talk to each other and we can expect vehicle interfacing as well. Rival consortia continue to steam ahead with their own ideas about what should control tomorrow's Internet of Things. Samsung offers the Artik platform (**https://www.artik.io**), a range of low-power IoT modules in form factors as little as 12mm square. The Allseen Alliance (**https://allseenalliance.org**) includes big names such as Sharp, Canon, HTC, LG, Qualcomm, Microsoft, Bosch, TP Link, and many more who are working on the open source Alljoyn operating system. Rivalling this is the Open Internet Consortium (**http://openinterconnect.org**) which counts Intel, GE, Cisco and others among its membership. With so many factions competing for supremacy, the quest for a *de facto* IoT protocol looks to be in turmoil.

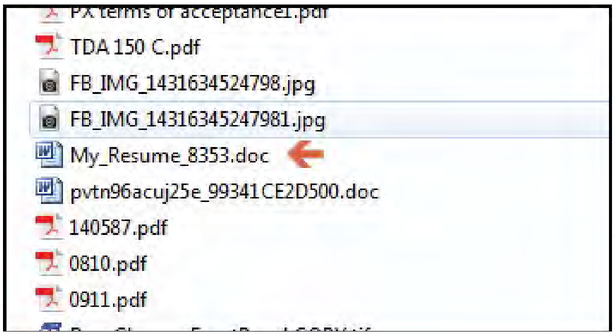
Deadly phishing

Following my item in June's issue about Secure Certificates and disabling SSL in the web browser (choose TLS instead), regular reader **Godfrey Manning** enquired about his choice of Kaspersky AntiVirus running on his PC. The popular and powerful AV software was suddenly displaying an error 'Unable to set up SSL connection'. Despite trawling the web for solutions and trying all manner of software settings in Windows, and scanning with free Malwarebytes Anti-Malware (fetch from **malwarebytes.org** only) the error message persisted and eventually I concluded that the problem must be at Kaspersky's end, with Godfrey's system unable to connect to their servers to update itself. It was eventually decided to install Avast Anti Virus instead.

The universal adoption of the Internet has brought with it the need to be constantly vigilant. Who would have thought that criminals halfway round the world would try to dip into your bank account at home? One of the jobs of *Net Work* is to remind users of some risks they currently face so that they know what to look out for and can spread the word, too. Criminals have become far more sophisticated with online attacks, and one of their main objectives is to slip a Trojan or malware onto your system without detection. Some attacks will faithfully record keystrokes or logins and send them back to thieves or blackmailers, or as I mentioned in the June issue, they may drop ransomware onto hard disks that encrypt essential data. The crooks then blackmail users by demanding Bitcoins to unlock the data once again. The damaged PC that I mentioned in June's issue was infected with Cryptowall 3.0 and I have since had a chance to check out its hard drive in my worklab: the malware had done an exceptional job of locking up every folder and encoding every data file. In the root of every directory they also deposited a ransom note in .html, .png and .txt format, also dropping it on the Windows desktop just to be sure. This would explain why the PC users were complaining that the machine was so slow, grinding to a near halt over several weeks: it was encrypting every file in the background until its job was done. Sadly, the old XP system had been 'protected' by an old version of AVG only. The machine itself was a write-off and last time I looked, a new PC was humming away in its place and business was bustling once again.

Apart from visiting infected websites, viruses are introduced via spam messages containing dodgy weblinks or carrying suspicious file attachments. At the time of writing, I happen to be awaiting an email from HSBC Bank, and sure enough a phony email just arrived in the guise of HSBC with the message starting 'Dear alan@epemag.demon.co.uk' – a sure sign of trickery, but spelling mistakes, poor grammar or generally strange

dialect are also dead giveaways. A spearphishing attack is a highly personalised spam mail, perhaps using the address and style stolen from one of your trusted correspondents to make it appear familiar and authentic. It is just too easy to click a malicious link or open a file when you're working flat out, and a careless mouse-click is all it takes to risk an attack, especially if you mean to right-click and shred it, but double-click it and run it instead. At such times you hope your AV does its job.



It is all too easy to accidentally double-click on an innocent-looking filename...
... but hopefully antivirus software will stop any threat.

Recently, another Trojan was discovered with the potential to deliver a devastating payload: unlike the Cryptowall-infected PC just mentioned (which was written off), the **Rombertik** malware will do the job for you by wrecking the hard disk's master boot record (MBR). Rombertik is a highly complex attack that arrives as a deceptively small spam mail attachment disguised as a Windows screensaver (.scr). We have all seen them, and Rombertik does a timely job of reminding us of the possible risks of opening suspicious files. Rombertik will endeavour to see if it is being caught and analysed and if so, it will try to trash the MBR and reboot the PC, wrecking the system. More details can be found on Cisco's blog at: **http://blogs.cisco.com/security/talos/rombertik**. They say that poison comes in little bottles, and innocent screensaver files like these can inflict deadly damage if users are caught unawares.

EPE online update

Last, a brief reminder about *EPE's* own online presence at: **www.epemag.com**. Each month's magazine has a dedicated webpage containing links to monthly downloads (source code, images, etc, which are rolled up into a single .zip file) along with a brief description of the issue's contents, updates, photos and more. A powerful new search facility is being added and visitors will be soon able to enter keywords and pinpoint an issue far more accurately than ever before. That's all for this month's *Net Work*. You can email the writer at: **alan@epemag.demon.co.uk**

By Robert Penfold

Pi transistor checker

IT does not seem that long ago that the B+ version of the Raspberry Pi was introduced, but this has now been superseded by the Raspberry Pi 2 Model B (Fig.1). This is largely compatible with the B+ model, with the same ports being present, and the B+ board layout being retained. Physically, there is no obvious difference between the two, as can be seen from Fig.2, and the new version is compatible with the cases made for the B+ model. The main chip on the new version is slightly larger, but that is about it. The only obvious lack of compatibility is that the Raspberry Pi 2 requires the latest version of the Raspian operating system, and it will not work with versions intended for the earlier models.

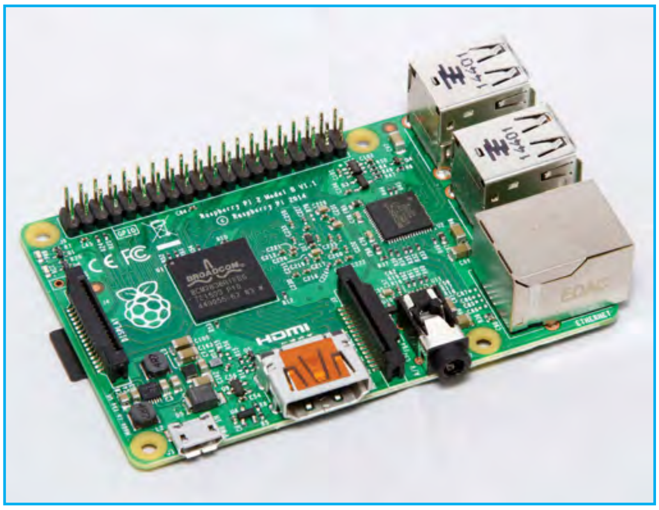


Fig.1. The Raspberry Pi 2 has the same ports as the B+ version, including the all-important GPIO port. It has the larger (40-pin) version of the GPIO port, but apart from using a different connector, it is fully compatible with the original (26-pin) port

Core changes

The main change is that the Raspberry Pi 2 has a quad-core chip, and is said to be up to six-times faster than the earlier versions, which all have single-core processors. The model 2 has its processor running at 900MHz, which compares to 700MHz for all the earlier versions. The increase in speed is reliant on the particular software being run, and its ability to exploit the quad cores. Depending on how well they are utilised, the boost in speed can be anything from a modest increase to the full six-fold improvement. In general, the model 2 does seem to be noticeably quicker, with the user spending less time waiting for something to happen when things stutter slightly.

There seems to be no problem in using the Raspberry Pi 2 with the circuits featured in this series of articles in the recent past. It runs Python 3, and has the same GPIO port as the Raspberry Pi B+. This version of the GPIO port has a 40 way connector with more input/output lines than the earlier type, but it is fully compatible with circuits designed for the original (26 pin) GPIO port. Just ignore the additional fourteen pins if you do not need them.

Old designs

With the ample supply of input/output lines on the GPIO port, it should be possible to use any Raspberry Pi with old

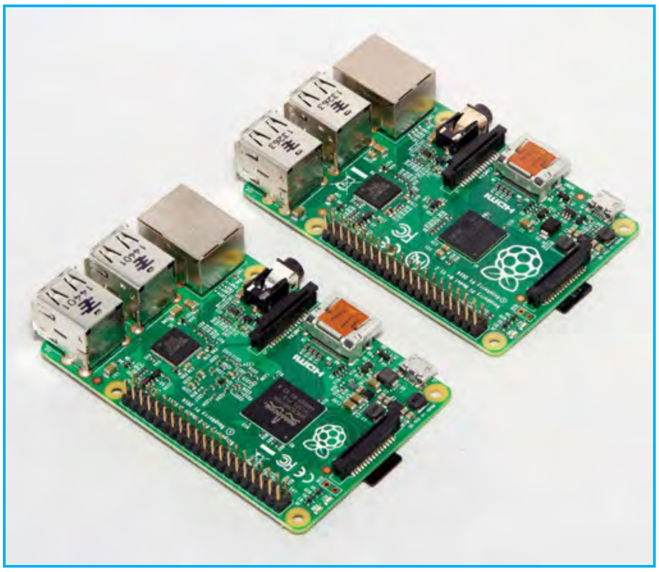


Fig.2. There is no obvious difference between the Raspberry Pi 2 (front) and the B+ (rear). The Raspberry Pi 2 has a quad core processor that makes it up to six-times faster than the earlier single-core types

add-on circuits designed for use with other parallel ports, such as the old PC printer port or even the User Port of the now 'antique' BBC model B. There are a couple of important provisos here, and the most obvious is that the components must still be available if you are building the add-on from scratch. This is clearly not a problem if you have a built-and-working unit, but no longer have a suitable computer port to suit it. It is then just a matter of trying to get suitable connections to the GPIO port sorted out, and writing a suitable control program.

The second proviso is that the add-on must not use a special facility that was available on the computer port originally used with the design, but cannot be provided by the Raspberry Pi's GPIO port. In most cases it would be possible to modify the add-on's hardware to provide facilities such as crystal-controlled clock signals and negative supply rails, but this would entail a substantial redesign and might not be worth the effort. I suppose it is possible that the original software might require facilities that could not easily be implemented using Python, although in most cases it would probably be possible to produce a simplified but fully usable Python alternative.

Transistor tester

As an exercise in converting an old design to suit the Raspberry Pi, I produced a suitably modified version of a PC-based transistor tester. The original circuit is shown in Fig.3, and it is designed to operate in conjunction with the printer port of a PC. Of course, these days it is a 'Ford choice' of USB or USB, and the parallel printer port is obsolete. Due to the lack of any supply outputs on a PC printer port, the transistor tester requires a 5V power supply.

The design is based on an AD557JN digital-to-analogue converter chip, which is one that has been used with the Raspberry Pi in previous *Interface* articles. On the face of it,

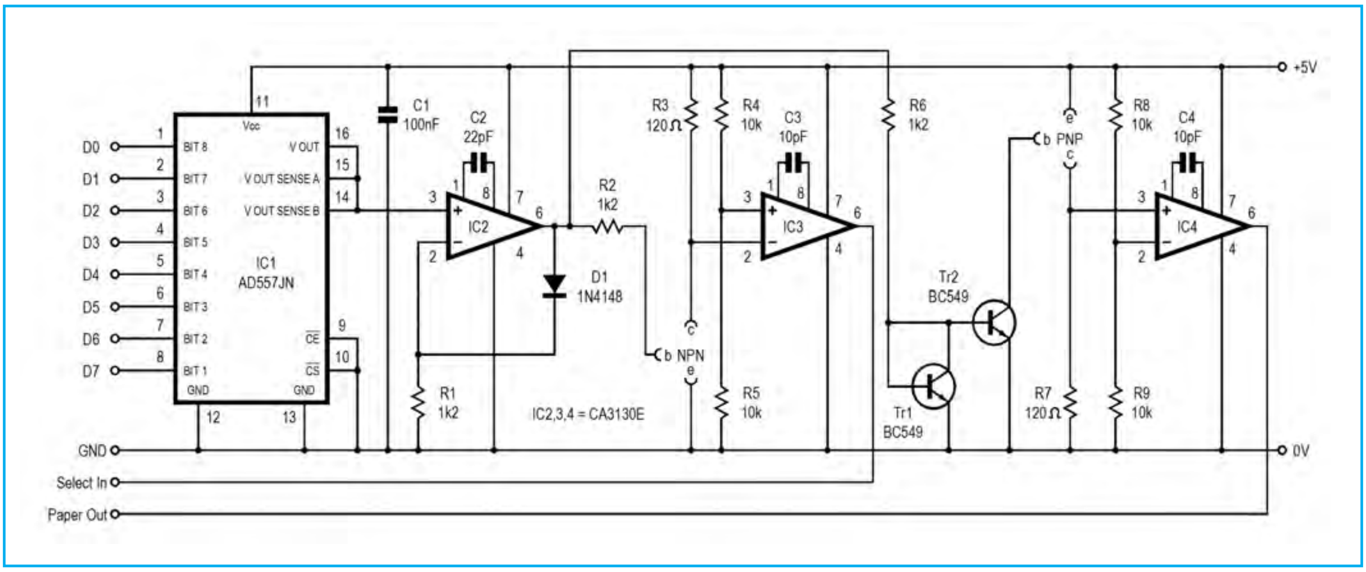


Fig.3. This transistor checker circuit is designed for use with the parallel printer port of a PC. Unfortunately, these days there are few PCs that have this type of port

this should make conversion of the circuit to operate with a Raspberry Pi relatively straightforward. Although the AD-557JN operates from a 5V supply, in practice its data inputs operate reliably from the 3.3V outputs of the GPIO port, and level-shifting circuits are not required. Driving the GPIO inputs from the 5V outputs of the transistor tester would be more dubious though, and it would be preferable to include level shifting here.

Fig.4 shows the Raspberry Pi version of the transistor checker. The only major change from the original circuit is on the output side where the printer port version had the NPN and PNP sections of the circuit driving separate inputs of the printer port. The Raspberry Pi version has the two outputs driving a simple NOR gate based on TR3, and this feeds a single input of the GPIO port. The NOR gate also acts as a level shifter that, with the aid of the 3.3V supply available on the GPIO port, provides an output signal at 3.3V logic levels. The GPIO port has a +5V supply line, and this is used to power the main circuit, thus avoiding the need for the separate supply required by the original circuit.

Counting up

Most transistor testers work by feeding a certain current to the base of the test device, and measuring the collector current flow. The current gain is equal to the collector current divided by the base current, and it is therefore proportional to the collector current. With a little mathematics the collector current can therefore be converted into the corresponding figure for current gain.

A potential drawback of this method is that low-gain devices produce a low collector current, and transistors tend to have relatively low current gains unless the collector current is reasonably high. This usually gives unrealistically low readings for low gain devices, and could even give the impression that these devices are completely dud. The method used by this tester is different, and it works by gradually incrementing the base current until the collector current exceeds 20mA. This ensures that there is a level playing field, with high- and low-gain devices all be tested at around the same collector current. The computer simply has to divide the 20mA collector current by the final base current in order to calculate the approximate current gain of the test device.

IC1 is the AD557JN digital-to-analogue converter, and it provides a maximum output voltage of 2.55V. IC2 acts as a buffer amplifier at the output of IC1, but the inclusion of D1 in the negative feedback loop produces an output potential from IC2 that is about 0.6V higher than the output voltage of IC1. This is done to counteract the potential of about 0.6V needed before a silicon transistor will pass a significant base

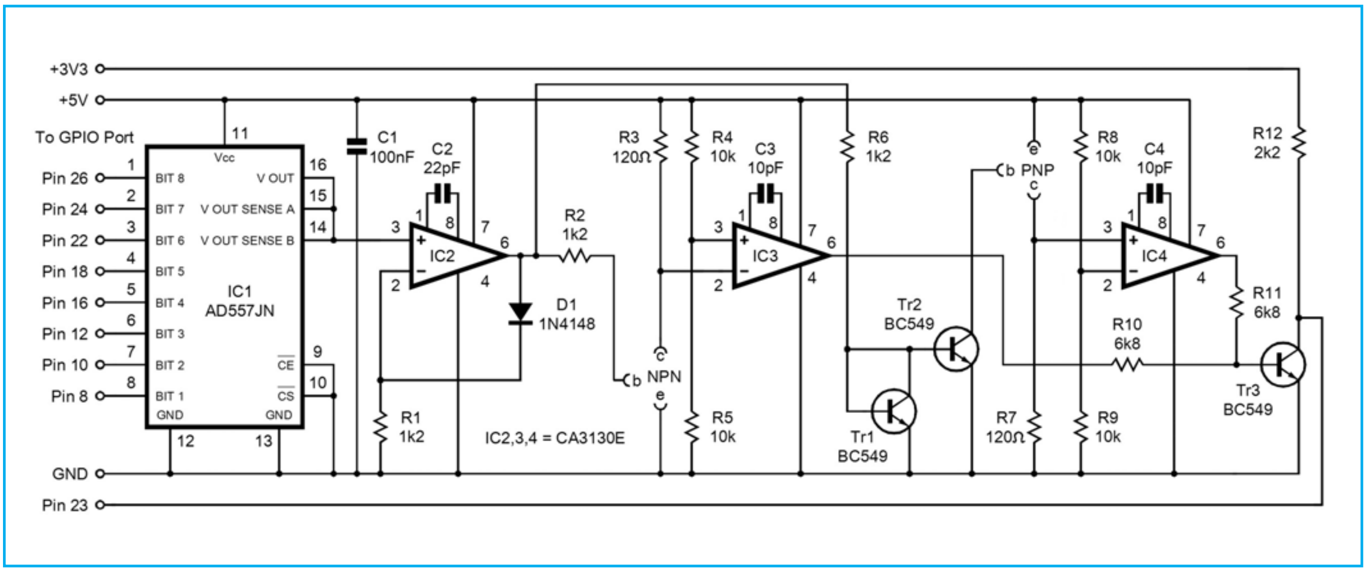


Fig.4. The Raspberry Pi version of the transistor checker circuit. The main change is that it has a NOR gate at the output, and it drives a single input of the computer port

current. The value selected for R2 sets the base current for NPN test devices at increments of about 8µA. R8 provides the same function for PNP test components, with the simple current mirror provided by Tr1 and Tr2 giving the required change in polarity.

Resistor R3 acts as the collector load for NPN test transistors, and the value of R3 gives a collector current of about 20mA with the collector at the mid-supply point. IC3 is an operational amplifier, but it is used open loop here, and it acts as a voltage comparator. The output of IC3 is normally low, but with a collector current of more than about 20mA its output goes high and switches on Tr3. This sends the collector of Tr3 low, which in turn takes pin 23 of the GPIO port low as well. This is detected by the software, which then halts the count to IC1, and calculates the gain of the test component. Things operate in a similar fashion for PNP test devices, with IC4 acting as the voltage detector. The roles of the two inputs of IC4 are reversed though, to allow for the fact that the collector voltage increases rather than decreases as the base current is ramped up.

Components

Any reasonably high-gain silicon NPN transistors can be used for Tr1 to Tr3, and any general-purpose silicon diode will suffice for D1. The requirements for IC2 to IC4 are much more stringent though, and most operational amplifiers will not work in this circuit. Only devices that can operate from a single supply rail are suitable, and a further requirement is that efficient operation must be obtained at a supply potential of just 5V. The CA3130E meets these requirements, but it is not internally compensated and requires external compensation capacitors (C2 to C4). The TS271CN also works well, and it does not require the external compensation capacitors. However, a suitable operating current has to be set by having a 1.2kΩ resistor connected from pin 9 of each chip to the 0V supply rail. All the integrated circuits are MOS types and require the standard anti-static handling precautions.

Software

A basic Python 3 program for use with the Raspberry Pi transistor checker is provided in Listing 1. This is based on the program for an AD557JN converter that was featured in a previous *Interface* article. Everything is set to suitable starting conditions by the initial section of the program. The next section is a `while… loop` that increments the value written to the converter chip by one on each loop, starting from zero. This normally loops until it detects that pin 23 of the GPIO port has been taken low. A variable called `mybyte2` is used to store the current count.

The next section of the program then increments the value in `mybyte2` by one if it is still at zero. This is just a simple way of avoiding an error caused by a subsequent program line dividing by zero. A count of zero indicates that the test device is placing a short circuit, or at least a very low resistance between the collector and emitter test sockets. A warning to this effect is displayed on the screen. The `while… loop` is terminated if the value in `mybyte2` reaches a value of 256, which is beyond the 255 limit of the 8-bit converter chip. If this occurs, the test device either has an extremely low current gain figure, or it is a dud that does not conduct between its collector and emitter to a significant degree. The final section of the program prints a suitable warning message if the count goes out of range.

The current gain of the test device could be calculated by multiplying the value in `mybyte2` by eight in order to give the base current in microamps (µA), and then dividing the 20000µA collector current by this figure. The program uses a rationalised version of this calculation, and simply divides 2500 by the value in `mybyte2`. This value is then printed on the screen, but it is limited to one decimal place so that long and unhelpful strings of figures after the decimal point are avoided.

It has to be admitted that this checker will not provide highly accurate results, and the resolution is much better at the low end of the measuring range than it is when testing high-gain devices. However, the tolerances on the current gains of most transistors are so vast that a ball-park figure is all that is needed in order to check them. The accuracy could be optimised by tweaking the values of R2 and R6. This tester is only suitable for NPN and PNP bipolar transistors, and it cannot be used to check any form of field-effect type devices.

Listing 1

```
import RPi.GPIO as GPIO
GPIO.setmode(GPIO.BOARD)
GPIO.setwarnings(False)
GPIO.setup(8, GPIO.OUT)
GPIO.setup(10, GPIO.OUT)
GPIO.setup(12, GPIO.OUT)
GPIO.setup(16, GPIO.OUT)
GPIO.setup(18, GPIO.OUT)
GPIO.setup(22, GPIO.OUT)
GPIO.setup(24, GPIO.OUT)
GPIO.setup(26, GPIO.OUT)
GPIO.setup(23, GPIO.IN)
GPIO.output(8, GPIO.LOW)
GPIO.output(10, GPIO.LOW)
GPIO.output(12, GPIO.LOW)
GPIO.output(16, GPIO.LOW)
GPIO.output(18, GPIO.LOW)
GPIO.output(22, GPIO.LOW)
GPIO.output(24, GPIO.LOW)
GPIO.output(26, GPIO.LOW)
mybyte2 = 0
loops = 0

while (GPIO.input(23) > 0) and (loops <
256):
    GPIO.output(26, 0)
    if mybyte2 & 1:
        GPIO.output(26, 1)
    GPIO.output(24, 0)
    if mybyte2 & 2:
        GPIO.output(24, 1)
    GPIO.output (22, 0)
    if mybyte2 & 4:
        GPIO.output(22, 1)
    GPIO.output(18, 0)
    if mybyte2 & 8:
        GPIO.output(18, 1)
    GPIO.output(16, 0)
    if mybyte2 & 16:
        GPIO.output(16, 1)
    GPIO.output(12, 0)
    if mybyte2 & 32:
        GPIO.output(12, 1)
    GPIO.output(10, 0)
    if mybyte2 & 64:
        GPIO.output(10, 1)
    GPIO.output(8, 0)
    if mybyte2 & 128:
        GPIO.output(8, 1)
    mybyte2 = mybyte2+1
    loops = loops + 1

if mybyte2 == 0:
    mybyte2 = mybyte2 + 1
    print ("Test Device Closed Circuit")
if mybyte2 > 255:
    print ("Test Device Open Circuit")
mybyte2 = (2500/mybyte2)
print ("%.1f" % mybyte2)
print ("Finished")
```

PIC n' Mix

Mike Hibbett

Our periodic column for PIC programming enlightenment

Revisiting Fritzing

WE'VE decided to take a break from the DIY oscilloscope this month, to return to an earlier promise – to take a look at making your own parts for use with Fritzing, the hobbyist circuit and PCB design package.

We covered Fritzing back in January. It's a wonderful program, designed specifically for hobbyists and makers. It's completely free, and can be downloaded from their support website at: **http://fritzing.org**

Our main reason for coming back to Fritzing is to start using it for documenting the design of *Pic n' Mix* article circuits; it looks like an ideal application for simplifying the task of drawing the wiring diagrams for circuits constructed on breadboards. We've become fond of these breadboards for prototype builds, and the availability of small, cheap circuits that are compatible with breadboards is making them an essential component in every hobbyist's kit box. Showing designs constructed on breadboards is difficult – a photograph always misses some detail or other. Fritzing makes images that are clear, easy to manipulate and beautiful to look at. As the images are rendered in SVG – a scalable vector graphics file format – if something is not easy to see on the screen then it's no problem, you can zoom in and the image is redrawn at a larger scale without loss of clarity or blurring. We love the SVG image format. However, it's not going to replace photographs. Photography still has it's place to show the physical realisation of a design, in all its practical limitations, so don't expect Fritzing to replace that any time soon.

The objective this month is to create a Fritzing version of our PIC development board, the LPLC. For that, we are going to need the LCD module, our LPLC board, a trimmer resistor and some hook-up wires. We concentrate this month on the LPLC board.

Finding new parts

A quick search through the list of available parts reveals a suitable trimmer, but no LCD. The LPLC board is missing too, naturally, but that's what we are going to make today.

The LCD module we are using is a popular part, so we thought we'd search the web to see if anyone else had created a Fritzing part, and sure enough, a simple Google for 'Fritzing TFT LCD part' brought up a library of Adafruit symbols. Fritzing parts are stored in files with the extension **.fzpz**, and we quickly spotted the file **2.2 TFT with MicroSD Breakout.fzpz** available for download.

With the file downloaded, we went looking for an 'import' function to copy the data into our parts database. This

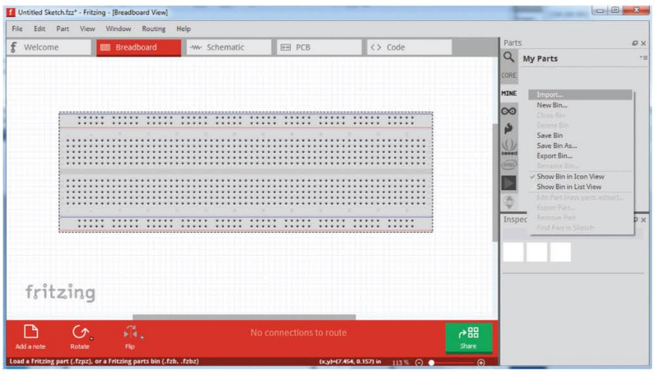


Fig.1. Importing parts in Fritzing

function is accessed by clicking on the MINE bin in the right-hand Parts dialog window, then right clicking in the empty area. This brings up the menu shown in Fig.1.

Clicking on Import and selecting the file you downloaded will copy the file into your parts bin. Select 'Save Bin' to confirm the addition.

With the Breadboard view visible in the left-hand window, it's a simple task of dragging the image of the part onto the view to add it to your Fritzing sketch. Unfortunately, in this case the part is not an exact match for our display because the pinout order is different, and there is an extra control signal – but we can modify it later to suit our needs. For now, clicking on the LCD in the Breadboard view and then pressing the delete key will remove it from the sketch.

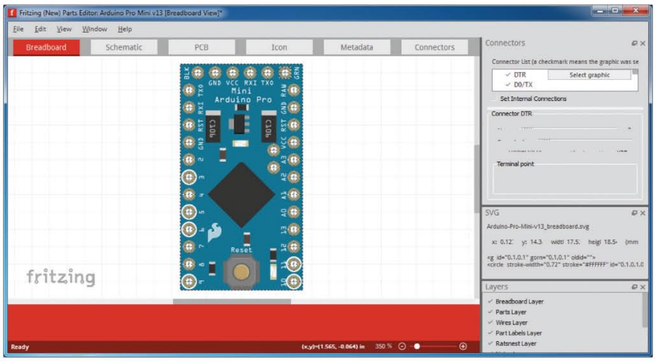


Fig.2. The Parts Editor

Making a new part

Now let's try and create a new part of our LPLC board. First things first, we hunt for a tutorial on part creation. This is normally the most complex task in any CAD program, and someone, somewhere, will have done a tutorial. Sure enough, Fritzing provides links to several. There is a catch though: SVG images are complex, and SVG image editors more so. The explanations for part creation are clear, but long. You are welcome to try, but we will take the lazy route – modify an existing part that is close to our desired part.

We start by finding an existing part that is close to our specifications. A look through the multitude of parts reveals the Arduino Pro Mini board in the Arduino parts bin. Dragging that onto the Breadboard view, you can then right click on it and select Edit. This brings up the (New) Parts Editor dialog window, shown in Fig.2.

Now things start to get complicated. First, there are four different sets of images and metadata to configure. Then, the hard part – images cannot be edited inside Fritzing, but must be edited using an SVG image editor. It's time to download Inkscape, the free editor, and learn how to use that. At the same time, we download the font that Fritzing insist is used in part designs, OCRA. It's an unusual font, originally designed for use with optical character recognition systems (it's used at the bottom of bank cheques for account identification.) It's free to download, but once again, this is something new. 'How do I install a font in Microsoft Windows?' Fortunately it's not terribly difficult – you open Control Panel and then select Fonts. This displays an Explorer-like list of files, and you simply unzip the downloaded font collection and drag the ttf files across. Inkscape will automatically register the new font the next time you start the program. The links to both downloads are listed at the end of the article.

We started off gently by modifying the metadata – the description of the part – in the Parts Editor. Then, selecting File > SaveAsNewPart, created a copy of the original part in our 'MINE' parts bin. The part was correctly labelled and the text reflected our changes. So far, so good.

Now comes the confusing part – how does the graphic image link to the schematic? When you hover the mouse over the pads in the image, the pad highlights. Yet there are no options in the Parts Editor to change these. Confusing.

SVG Graphics

The answer lies in the use of SVG graphics images, and the need to install a special image editor. SVG images are not images at all, but text files that *describe* the image. You can open one in notepad and take a look; the file format is xml, a standard file format used to describe webpages on the Internet.

A close look at the PCB file for the part we are using, **Arduino-Pro-Mini-v13_pcb.svg**, reveals how the pads are identified. Close to the bottom of the file is the line:

```
<circle id="connector14pad" fill="none" stroke="#F7BD13"
stroke-width="1.224" cx="43.2" cy="3.6" r="2.052"/>
```

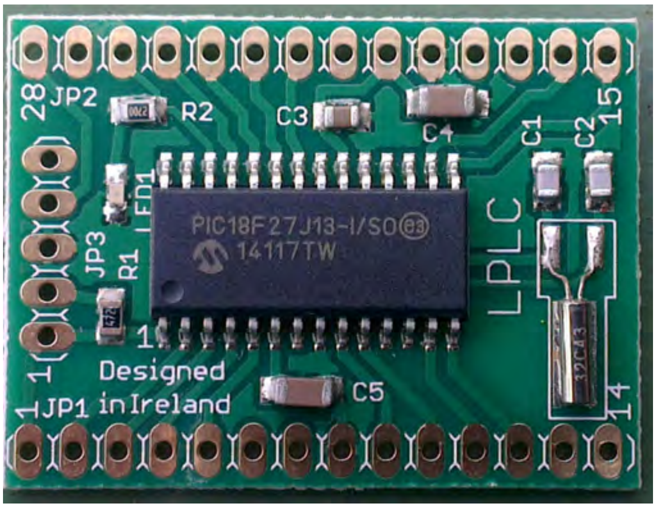


Fig.4. The LPLC Board

Fritzing relies on entries like these to be present in the SVG files, and uses them to identify the connector pad graphic image sections. This is quite neat; although it makes our life more complicated, it simplifies the design of Fritzing. The complexity of creating and editing graphics is left to third-party SVG editors that are probably better suited to the task.

Armed with this knowledge, we can now take a look at the **.svg** files currently in use by our new part. In the Parts Editor we can see that for each of the views – Breadboard, PCB and Schematic – the three different **.svg** files being used. We start by opening the Breadboard image in InkScape. The image consists of groups of items (which will be very important when we get to the PCB view), and you use the mouse right-click menus to navigate into a group, and double click with the left to select individual items. We quickly navigated to one of the pads, and sure enough in its Object Properties field the Label is '#connector25pin'. It's this identity that Fritzing uses to locate the actual sub-image within the file for the connector location. You can see this in Fig.3.

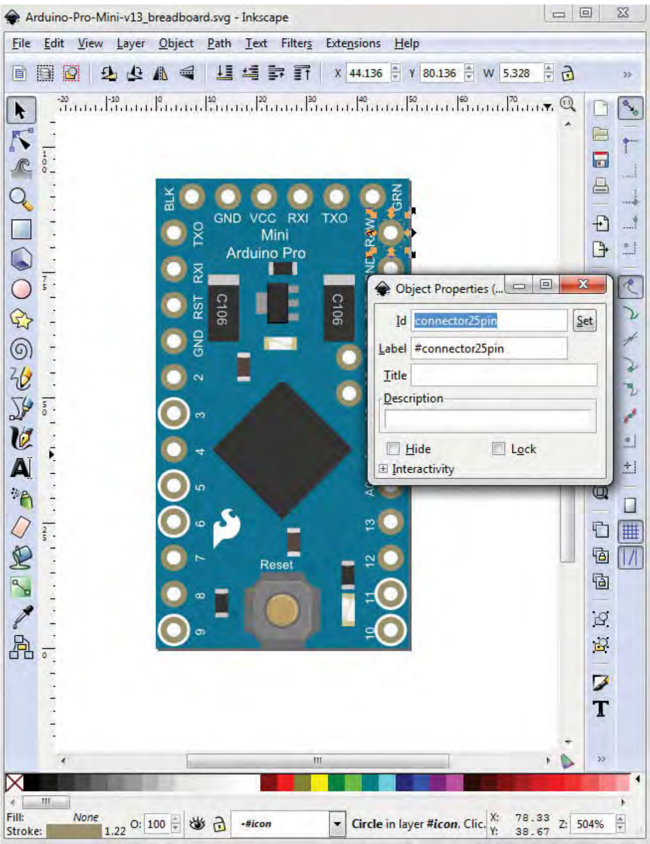


Fig.3. Editing an SVG image in Inkscape

We can now edit the drawing, moving and copying pads and creating new shapes for additional text and components. The components in this image are cosmetic only – they do not affect any part of Fritzing's operation. The pads, however, do. So we have to be careful with placement (0.1-inch alignment) and naming.

Our LPLC board is shown in Fig.4. As this board is our own design we took all dimensional data from the CAD package, but in most cases you will have to refer to the manufacturer's datasheet, or use a calliper to measure hole diameters and distances. A ruler will do at a pinch, but remember that some of these images are used by the PCB design tool in Fritzing, and so your measurements of pad spacing and hole diameters must be accurate.

We start by opening up the existing part in the Parts Editor. We will save our changes as a new part, leaving this one unchanged. First, we click on the 'Connectors' tab in the Parts Editor and delete the old connectors (setting the 'number of connectors' to 1, then deleting that one with the 'X' button.) Then we enter a value of 28 for the number of connectors. We don't add the programming interface header pins, as you won't need to wire these to anything – the PicKit3 programmer plugs in there.

The connectors are all uniquely identified from pin 1 to pin 28, but we can also add the part-specific name for these pins in the description field.

Next, we select the Breadboard view, and select File > Load image for view. Navigate to where you stored your **.svg** file and select it. At this point you may get a warning about fonts if you have not used the OCRA font favoured by Fritzing – the fonts will be automatically converted if you have not.

Creating the PCB image was easier – copy the existing part's **_pcb.svg** file, and modify that. The PCB data for the LPLC board is just two strips of 0.1-inch pads separated by 0.9 inches, so it was a straightforward task. Repeating for the schematic view we do the same; this is a simple line drawing of the logical layout of the connections to the LPLC. On all three views you have to select the graphics image for each pin.

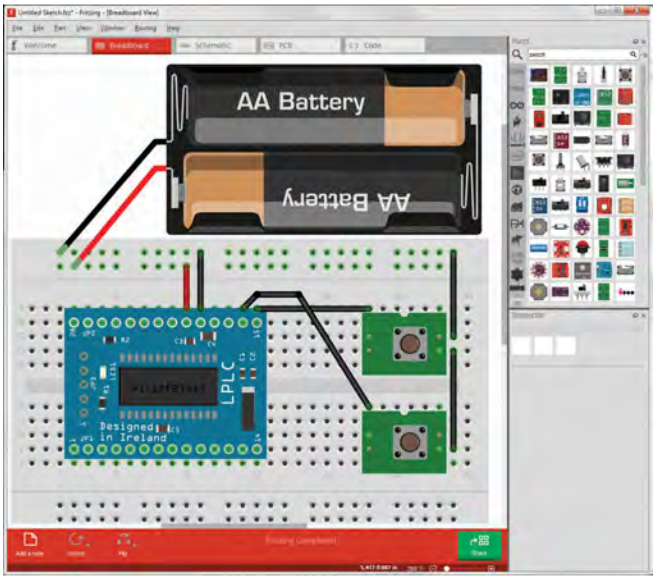
The Icon tab is used to specify an image that will be displayed in the Parts bins, where you choose components to use in your sketch. You can use a menu option to select one of the other images already used (for the breadboard, PCB or even schematic) or you can supply a completely different image.

Once you have connected all the pins, simply select File > Save as new part, entering the prefix 'LPLC'.

Phew! That was hard work, but we learned a lot along the way. You can now drag the part onto a breadboard view and start wiring it up just like any other part. A very quickly drawn random assembly of parts can be seen in Fig. 5.

Conclusion

Hopefully you can see that although creating parts is far from trivial, it's not an enormously complex task, and the results are certainly worth it. SVG images are far more versatile than



simple bitmap images, and it has us thinking of how we might use the file format in our own programs. The LPLC part that we have created is available freely online, so you won't need to go through the same pain – simply download from the magazine website at the usual location, on the page for this month's issue.

Next month

We return to the oscilloscope project next month, and you can expect a glorious Fritzing image to go with it!

Not all of Mike's technology tinkering and discussion makes it to print. You can follow the rest of it on Twitter at ***@MikeHibbett****, and from his blog at* ***mjhdesigns.com***

References

Ref. 1 OCR-A font, available from: **http://osdn.jp/projects/tsukurimashou/downloads/56948/ocr-0.2.zip**

Ref. 2 Inkscape, available from: **https://inkscape.org/en**

Fig.5. The Fritzing version of the LPLC board

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Noise – Part 1: Noise, distortion and spectra

THIS month, we will look at a topic suggested by *EPE* editor Matt Pulzer, that of noise in electronics. We usually cover topics raised in the *Chat Zone*, but occasionally look at other topics, which will help readers working on their own designs or working with technical documents from device manufacturers. Noise is important in circuit design, but is a complex subject with plenty of jargon to get to grips with. In this article we will look at some of the basic concepts and definitions related to noise, in particular with reference to the frequency spectra of signals. In a later article we look at noise analysis of circuits.

In circuit design the word 'noise' can be used to refer to any undesired signal that disrupts or obscures a wanted signal. However, the term is often used more specifically to refer to the random signal variations occurring inside the circuit or system – this noise is generated by the components themselves, by various mechanisms, which we will discuss in a later article. This distinguishes noise from externally induced perturbations, known as interference, and non-random (systematic) signal changes produced by the circuit itself, known as distortion. Random noise and distortion are sometimes considered together as they are both unwanted components of the output originating within the circuit.

The previous paragraph is perhaps over simplified. Interference can occur within a circuit or system; examples include crosstalk between multiple channels or signal paths, and digital signals being coupled into analogue sections of mixed circuits. Random noise generated outside a system can be picked up (as interference) and added to the noise from within the system.

Random

Radom noise causes the instantaneous value of a signal to deviate from its 'true' value, with decreasing probability for larger deviations. The specific mathematical function for the probability versus amount of deviation depends on the type of noise, but it may be the Gaussian or normal distribution (the 'bell curve', well known in statistics), in which case we have 'Gaussian noise'.

There are various types of random noise generated within electronic circuitry; these include thermal noise, shot noise, flicker noise, and avalanche noise. This generated noise is fundamentally due to the discrete nature of electricity at the atomic level – electric charge in circuits is carried in packets of fixed size via electrons or holes.

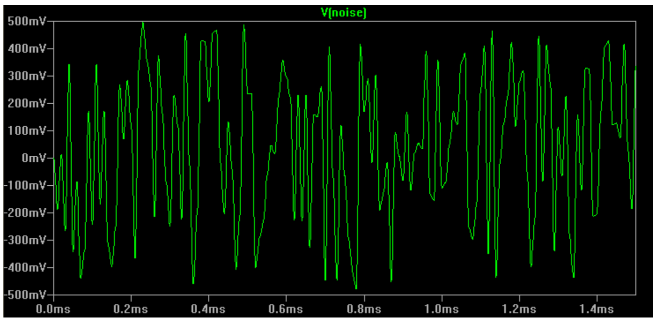

Fig.1. Random signal generated using LTSpice

In radio systems the antenna will receive random noise radiated by its environment. All objects at temperatures above absolute zero radiate electromagnetic energy, which may be picked up by antennas, or parts of any circuit behaving as an antenna. Sources of noise include the ground, the atmosphere, astronomical bodies, and even the cosmic background radiation from the origin of the universe. The combined noise from the atmosphere and extraterrestrial sources is referred to as 'sky noise'. This ambient electromagnetic radiation is very important in some communication systems, but in most circuits the internally generated noise (circuit noise) dominates.

The general usage of the terms 'noise' and 'interference' may not be very precise, however, we can be much more precise about our definitions when discussing particular types of both. In the case of noise, we are often specifically interested in random noise, which we have seen can be external (eg, sky noise) or internally generated by a circuit or system (circuit noise). In this article we will mainly concentrate on random circuit noise. The word 'random' indicates that the fundamental processes of random noise can be studied using statistical theory, and indeed this is the case, however, here will avoid use of advanced mathematics.

The waveform in Fig.1 shows a random voltage variation with time. This gives us a simple insight into what noise 'looks' like, but in general, plotting random noisy signals against time is not particularly useful. When dealing with noise we often need to look at the spectrum of the signal – the variation of signal level against frequency.

Unwanted signals can look like random noise (eg, on an oscilloscope), but actually have significantly different characteristics. For example, the noise on the power supply of a digital circuit may look random, but a look at the spectrum will show that certain frequencies, related to the system clocks will be dominant. The noise is caused by transient currents, which flow when gates switch. The gates do not all switch together because of varying delays in the circuit, and they do not all switch in the same cycle due to data variations, so there is some randomness; however, the switching is coordinated by the system clock(s) so there will much stronger components of the spectrum at the frequencies related to the clock(s).

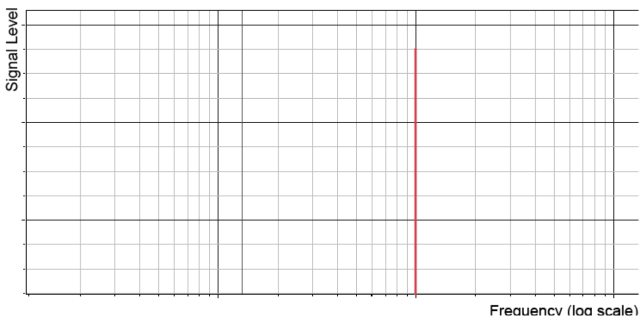

Fig.2. The spectrum of a pure sinewave has a single peak at the frequency of the sinewave

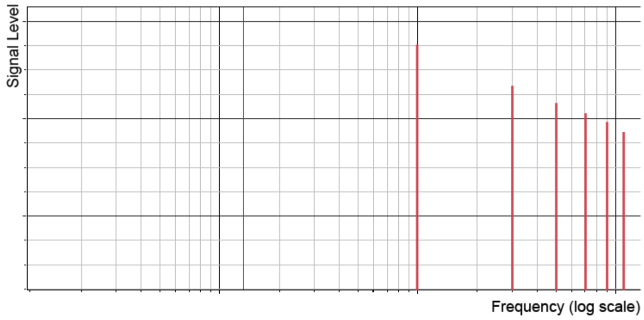

Fig.3. The spectrum of an ideal, simple periodic waveform such as a square or triangle wave has multiple individual peaks at the frequencies related to multiples of the waveform's period

Spectrum

The sinewave has the simplest spectrum, with a single peak as just one frequency (see Fig.2). Other simple periodic waveforms, such as square waves, have spectra with peaks at specific sets of individual frequencies (Fig.3). Complex, meaningful waveforms, such as voice signals, contain a wide range of different frequencies, but with stronger components at some frequencies than others and complex variation of signal strength with frequency (Fig.4). In contrast to all of these, random noise has a smooth continuous spectrum (Fig.5 and Fig.6).

Real signals will always have some noise, which will show up in any measured spectrum, unlike the noise-free cases in Fig.2 and Fig.3. The noise part of a real measured spectrum will tend not to be perfectly smooth, as shown in Fig.5 and Fig.6, although averaging the measurement over a long time span will tend to give a smoother result. Fig.7 show a more realistic version of the spectrum in Fig.3.

Distortion creates peaks in the output spectrum that were not present in the input, so Fig.3 could feasibly represent the output of a distorting, but noise-free, circuit with a sinewave input represented by Fig.2. Fig.7 could represent the output of a circuit that adds both noise and distortion to the ideal sine input represented by Fig.2.

The fact that every component in any electronic circuit or system generates random noise, and the rest of the universe is radiating noise, means that there is always a certain level of noise, even with no signal present – this is the 'noise floor'. In Fig.7, we see the noise floor as the low-level, almost constant values across the entire frequency range and the wanted signal as the peaks (assuming in this case they are all wanted) at specific frequencies and much higher signal levels. It follows that if signal levels diminish to levels at or below the noise floor they cannot easily be detected or measured (they 'disappear into the noise').

If the properties of the required signal are known (eg, by using special coding sequences in communications) then there are techniques that can extract signals that are even smaller than noise present within the signal.

The spectra of ideal periodic waveforms (Fig.2 and Fig.3), which are discrete lines at single frequencies and those of complex signals, which can be assumed to vary continuously with frequency, are fundamentally different. If you plot a single frequency point on a continuous graph it is infinitely small and hence invisible. Thus, for the line spectra (Fig.2 and Fig.3) we are actually plotting the signal level corresponding to a small but finite band of frequencies.

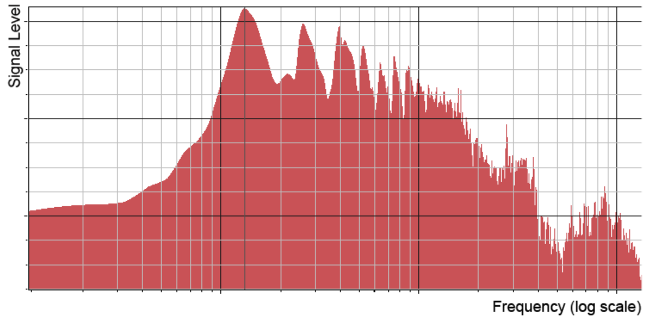

Fig.4. A complex waveform such as speech has a varied spectrum with complex changes in signal level at different frequencies

Building a spectrum from a set of frequency bands corresponds to what happens when we use a measuring instrument (spectrum analyser) to obtain the spectrum. The width of the bands corresponds to the frequency resolution of the instrument. Similarly, simulated or calculated spectra will have frequency bands corresponding to the resolution or detail level of the calculations performed.

Careful with that axis

Care must be taken when looking at spectra to note what axes are being used. The frequency axis may be linear or logarithmic (it is logarithmic in Fig.2 to Fig.7). Linear axes are more likely to be used when a small range of frequencies is being considered, and for periodic signals, such as the example in Fig.3, for which the peaks would be even spaced on a linear axis. A linear frequency axis may make it easier to see harmonics (multiples of the fundamental signal frequency).

The *y*-axis in Fig.2 to Fig.7 is somewhat ambiguously labelled 'signal level'. In general, this axis of a spectrum could be voltage (or current), or power – usually expressed voltage squared or current squared, which is proportional to power. We can use the square of voltage (or current) directly without knowing what resistance is involved, as the resistance is assumed to be constant and we often plot the spectrum relative to a reference power level, rather than as an absolute value. The signal level scale is often logarithmic, usually with decibel-based values (in which case reference level is definitely being used). For noise spectra, however, the plotted quantity is also likely to be power density, which may need some further explanation.

Random noise signals have an average voltage of zero, and unlike signals such as sinewaves, there is no clearly defined peak voltage, just a probability of being a particular voltage, as mentioned earlier. Therefore we need to use a power-based measure of noise level. This could be (the average of the) voltage squared, or the square root of voltage squared (as in the RMS (root-mean-squared) values commonly used for AC measurements).

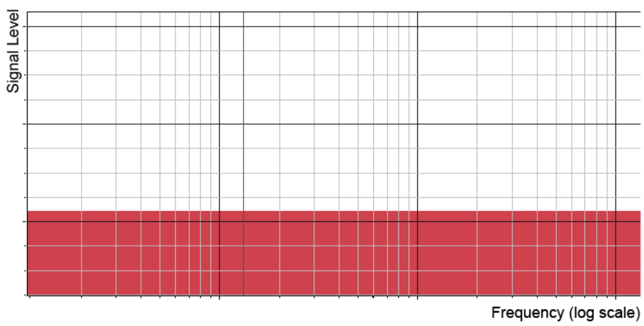

Fig.5. Radom noise has a smooth spectrum which may be constant, as shown here, or steadily changing

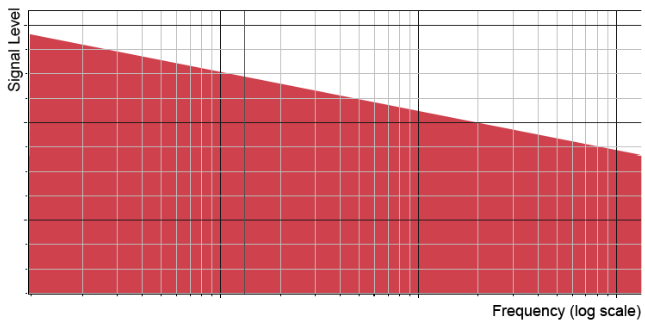

Fig.6. Radom noise has a smooth spectrum which may be constant, or steadily changing as shown here

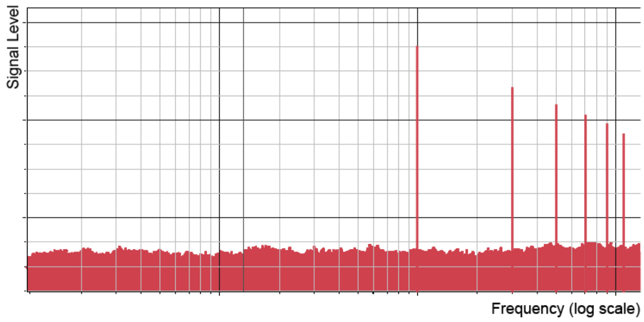

Fig.7. There is always random noise present in real signals, so you do not see real spectra like Fig.3, also measured noise will not have a perfectly smooth spectrum unless measurements are averaged for a long time

As already discussed, the spectrum is built up from a set of arbitrarily narrow frequency bands. It follows from this that if we are plotting power and we double the width of the band the power in each band approximately doubles (exactly doubles if the power is constant across the range). As it is the measurement, not the signal, that has changed here, it would be better to use a measure that is independent of the band size. Power density, or power spectral density (PSD), which is measured in watts per hertz (W/Hz) fulfils this requirement. We can also use the square root of this (corresponding to the RMS voltage), which means the signal level is spectral density, measured in volts per root hertz (V/Hz^{½} or V/√Hz). Use of spectral (power) density effectively normalises the frequency band used for the spectrum measurement to 1Hz (whatever band was actually used).

PSD is often expressed logarithmically in decibels. As PSD is a single value, rather than a ratio, the decibel value is found relative to a reference level, which is most commonly 1mW. The symbol dBm is then used for the decibel value. A power value in dBm is found using 10log(power/1mW). A PSD spectrum using dBm has y-axis units of dBm per hertz (dBm/Hz).

Any colour you like

Random noise may be classed according to the shape of its spectrum (eg, see the difference between Fig.5 and Fig.6). White noise has the same power throughout the frequency (f) spectrum, whereas $1/f$ noise (or pink noise) decreases in proportion to frequency. For pink noise, there is the same amount of noise power in the bandwidth of say 100Hz to 1kHz as there is in 1kHz to 10kHz, whereas for white noise there would be 10 times as much power in the bandwidth 1kHz to 10kHz as 100Hz to 1kHz because it is 10-times larger. Other noise colour terms are used, but are generally less well known. Red noise decreases in proportion to f^2. Blue noise increases in proportion to frequency and violet noise increases in proportion to f^2.

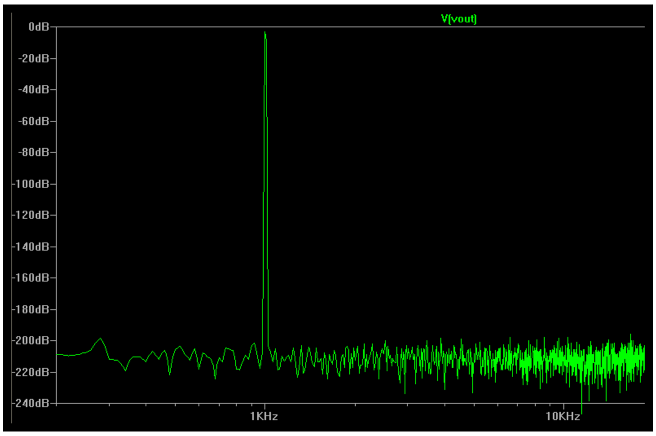

Fig.9. Spectrum of an ideal sinewave calculated from transient waveform data in LTSpice

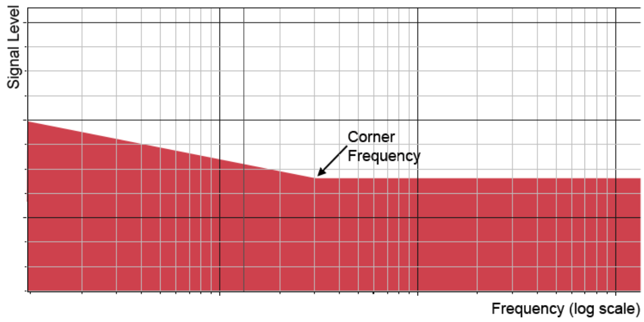

Fig.8. Typical spectrum of amplifier noise

If we plot the spectrum of white noise it will be flat, as in Fig.1. This is true, even though the frequency axis is logarithmic if the frequency bands used for the measurement of the spectrum are of a fixed size throughout the measured spectrum (the usual case). This means the bands are narrower *as drawn on the graph* at higher frequencies. Whether or not this is easily visible will depends on how the plot is created from the raw data. As the signal level is constant for white noise, the spectrum shape does not depend on whether the signal-level axis is linear or logarithmic. If we plot pink noise on a log/log axis (eg, dB level vs. log frequency), we get the spectrum shown in Fig.6 – a straight line decrease in level with frequency. The PSD falls at 10dB/decade. On a linear axis the spectrum is smoothly curved ($1/f$ shape).

Amplifiers (and other circuits) typically exhibit a mixture of pink and white noise, with pink noise dominating at low frequencies. The frequency at which the dominant noise component changes between pink and white noise is called the 'corner frequency' or 'noise corner' (see Fig.8).

We have described spectra as being made up from data for a set of frequency bands – this is like a histogram and leads to the 'filled in' format of the plots shown on Fig.2 to Fig.7. However, the spectrum does not have to be plotted like this – a line graph can also be used. An example is shown in Fig.9, which was obtained from a simulated ideal sinewave in LTSpice. Although this uses an ideal sinewave, the plot is more like Fig.7, with an obvious, albeit low, noise floor, that is like the ideal sine spectrum in Fig.2. This is because the inevitable numerical errors in the calculations (eg, rounding errors) behave like noise. The fact that the frequency bands get closer at high frequencies can be seen in Fig.9.

Noise metrics

The difference between the signal and the noise is often of great importance, this is expressed as the signal-to-noise ratio (SNR), usually in decibels (dB) and based on the ratio of noise power (hence the v² terms in the equation). Larger values indicate better performance.

$$SNR = \frac{p_s}{p_n} = \frac{v_s^2}{v_n^2}$$

$$SNR_{dB} = 10\log_{10}\left(\frac{v_s^2}{v_n^2}\right) = 20\log_{10}\left(\frac{v_s}{v_n}\right)$$

Where p_s is the signal power, p_n is the noise power, v_s is the RMS signal voltage and v_n is the RMS noise voltage. When using or quoting SNR values, the bandwidth (range of signal and noise frequencies considered) should be quoted because, as we have seen, noise power is frequency dependent and noise may be present well outside the range of signal frequencies of interest. Note that we use 10log(x) to express x in dB when x is a power ratio. If x is a voltage or current ratio we use 20log(x).

The noise factor, F_N, of a circuit is a measure of how much noise the circuit adds to the signal. Lower values indicate better performance, with a noiseless amplifier having a noise factor of 1.

$$F_N = \frac{SNR_{In}}{SNR_{Out}}$$

The similar sounding term, noise figure (NF), is the noise factor in decibels. An ideal (noiseless) amplifier has a noise figure of 0dB

$$NF = 10\log_{10}\left(F_N\right) = SNR_{In,dB} - SNR_{Out,dB}$$

Noise is not the only unwanted component of an output signal introduced by a non-ideal circuit. There is distortion that is due to non-linearities – this is obviously of importance in circuits which are supposed to be linear, such as amplifiers. In the simplest case, the input signal is a sinewave (ie, a single frequency in the spectrum) and distortion introduces additional frequencies at multiples of the input frequency. These additional frequencies can be seen in the spectrum of the output and are referred to as harmonic distortion. Unwanted (spurious) peaks in the spectrum are commonly called spurs, whatever their specific cause is.

The ratio of the sum of the RMS values of all distortion components to the RMS value of the wanted signal is called the 'total harmonic distortion' (THD). THD plus noise is sometimes specified and is written as THD+N. This is defined in a similar way to THD, but the RMS noise value is added along with the distortion components. THD+N is used because it is relatively straightforward to measure and gives an indication of the overall 'goodness' of the output. To measure THD+N the original signal frequency is removed from the output using a very narrow band filter (notch filter). The ratio of the measured input, to filtered output, RMS signal values is the THD+N.

Another common noise and distortion parameter is 'signal to noise and distortion' (SINAD) which is defined as:

$$SINAD = \frac{Signal}{Noise + Distortion}$$

Larger values indicate better performance. Similar to THD+N, SINAD can be measured using a sinewave input with the output notch filtered to remove the wanted signal and this level is compared with the unfiltered signal. In this case, the measurement of SINAD is:

$$SINAD = \frac{Signal + Noise + Distortion}{Noise + Distortion}$$

Both definitions can be found in various sources. For cases where the signal is relatively strong, the two values are close. SINAD is used to measure the sensitivity of radio receivers (ie, what signal is required to given an acceptable SINAD). For ADCs, SINAD provides a good indication of dynamic performance. Like SNR, SINAD is based on power and expressed in dB.

The 'spurious free dynamic range' (SFDR) is the ratio between the wanted signal (again assuming it is a sinewave) and the largest spur in the output signal, whatever the cause – the spur may be due to distortion, but it does not have to be.

When looking at noise metrics such as those just described always 'read the small print'. Some definitions do vary and it is always important to know the conditions, such as the frequency range for which the value is given.

AUDIO OUT

By Jake Rothman

RIAA equalisation – Part 2

Synthesising the curve

There are at least six different ways of generating the RIAA equalisation curve, along with the required 30 to 40dB gain. Most techniques use two capacitor sections in one network arranged in one feedback loop. The resulting section-to-section interactions make calculations difficult, but thankfully the mathematician SP Lipshitz solved this problem in a paper in the *Journal of the Audio Engineering Society* (June 1979). Also, the *National Semiconductor Audio/Radio Handbook* (1980) has some useful equations. This work was later refined into more usable optimised values by Douglas Self, who has been working on it 'forever', and his latest article in *Linear Audio* (Volume 7) offers more insights. Most of the time, values from these circuits can simply be scaled to suit one's available component values. If a designer reduces the capacitance by say 30%, then the resistance needs to be increased by the same amount. The standard 'Self circuit' is shown in Fig.10. The best networks, such as this one, can achieve accuracies of 0.01dB, which is over-engineering relative to the poor frequency response accuracy of ±2dB of the best cartridges. Douglas Self puts a compensating network to knock off the +0.38dB error at 20kHz due to the fact that a non-inverting circuit gain cannot fall below unity. I sometimes don't bother with this and use high-frequency roll-off elsewhere in the system, such as in the tone control circuit. It's essential to have bass and treble controls with vinyl, since correction is often required. A Baxandall or Ambler tilt control centred on 1kHz in the middle of the RIAA curve is ideal.

I remember seeing John Linsley Hood's RIAA pre-amp in *Hi-Fi News + RR* (Feb 1979) that used an inverting configuration to avoid the unity-gain problem and get a nicer looking square wave, but the 47kΩ input resistor made the noise eight-times worse than a non-inverting stage.

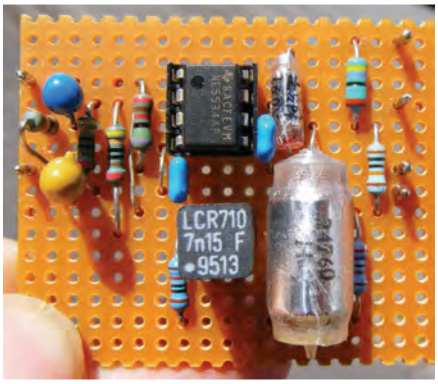

Fig.12. Veroboard version of scaled Self circuit. I've introduced many young people to vinyl with this cheap circuit

Since vinyl's frequency response is far from flat, most standard Hi-Fi designs use 5% tolerance components, the better quality British designs today use 1% metal-film resistors and 2.5% polypropylene capacitors. Some esoteric equipment uses 0.1% resistors and 1% polystyrene capacitors. My efforts to rescue perfectly good non-RoHS (Restriction of the Use of Certain Hazardous Substances) components from landfill has led me to acquire boxes of Philips lead-foil and Suflex/LCR aluminium foil polystyrene capacitors, which are the perfect audio EQ capacitor type. They have 1% tolerance and provide a complementary temperature coefficient with metal-film resistors. I sometimes use cheap (five pence) 1%

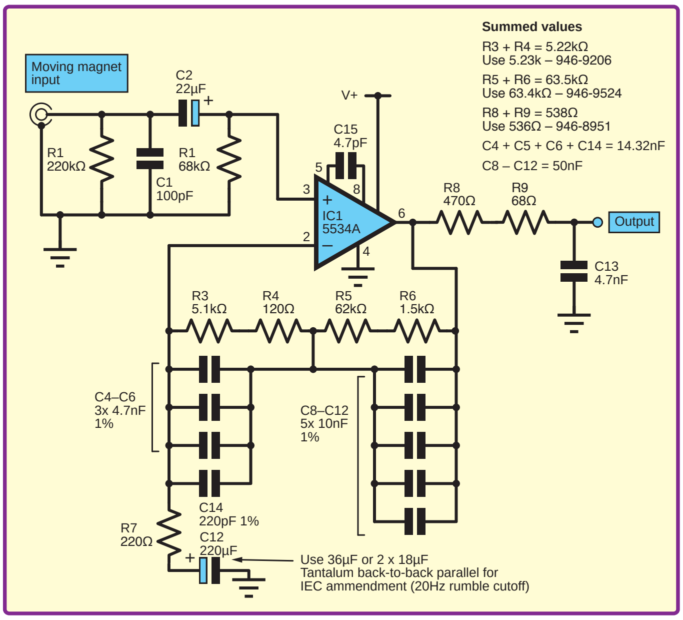

Fig.10. Douglas Self RIAA stage from his book Small Signal Audio Design. *This has become an industry standard with values scaled. Capacitors made from multiples of smaller values can have better tolerance assuming variations are truly random (in this case, the ideal total value is 50.15nF). If the maker has stripped out the accurate ones from the batch, as I have found with some encapsulated plastic cased types, they won't be. Old fashioned foil-wound axial capacitors do seem to exhibit true random variations (I've been playing with my Peak analyser!)*

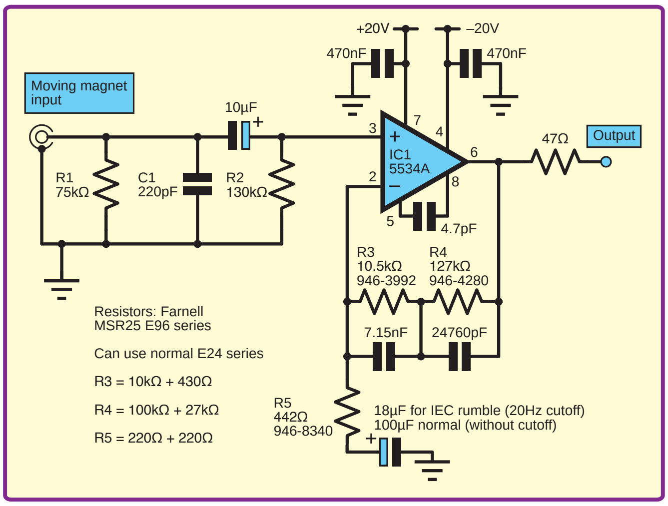

Fig.11. Scaled-value Self circuit – here the capacitors have be halved and the resistors doubled

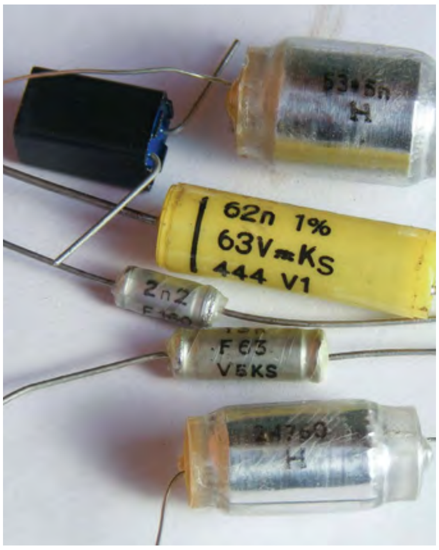

Fig.13. Rescued non-RoHS capacitors. Close-tolerance polystyrene types are no longer made. Lead foil is banned and the plastic film melts at too low a temperature for flow soldering. Polypropylene types are the next best. The only capacitor dielectric better for audio is Teflon

E96 Vishay MSR25 resistors from Farnell to avoid combinations of E24 types, since it pays with self-assembly to minimise the parts count. The Farnell part numbers are shown in the diagrams.

Fig.11 and Fig.12 show an RIAA preamp using these special surplus polystyrene capacitors. They were unusual custom values that just happened to scale to Self's design. I doubled the lower-arm resistor, which put up the noise by 1dB – this shows on the 'spec sheet', but you can't actually hear it with real records. (More of these capacitors are shown in Fig.13.)

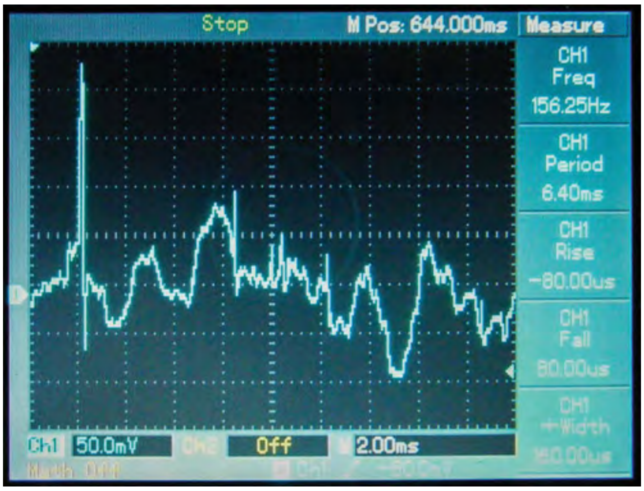

Fig.16. Record scratch displayed on a digital storage 'scope. These short, high-amplitude pulses can play havoc with amplifiers

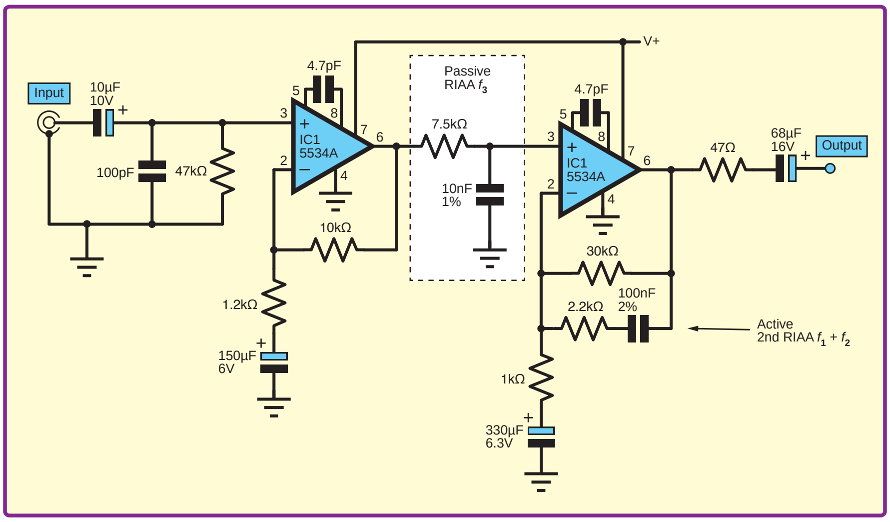

Fig.14. Two-stage RIAA preamplifier, lower noise but more prone to overload

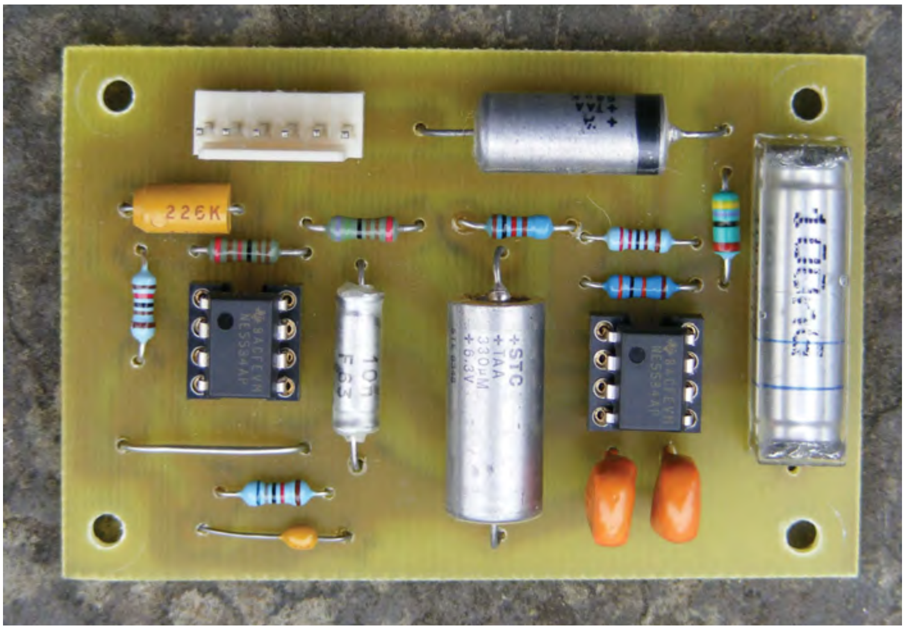

Fig.15. Two-stage RIAA PCB designed for the Modern Amateur Electronics Manual

Dual-amp RIAA amplifiers

In some RIAA amplifiers the two EQ sections are split into two, as shown in Fig.14 and Fig.15. This enables the much simpler $1/2\pi RC$ equation to be used with little interaction and often allows standard E6-value capacitors. Also, the very large op-amp-style open-loop gain requirement of the single-stage circuit (80dB) isn't needed and can be built with simple discrete stages. The noise is often reduced, but at the expense of overload capability.

Mechanical *vs* electronic noise

In practice, the noise is not very important since the record surface noise dominates that of almost any amplifier, by a factor of at least 10. Audio engineers worry unduly about phono amplifier noise, because it can be clearly heard when no record is playing when the volume control is turned up. However, once the stylus makes contact with the rotating record the electronic noise is completely swamped by mechanical noise. The overload capability is much more important, because the peak amplitude of the scratches can be 10dB above the music. Fig.16 shows a typical scratch. If clipping occurs, it sounds much worse than any noise because a listener can't hear though it. If the amplifier 'hangs up' that's even worse, since the music is blocked for a while. Some RIAA amplifiers use passive equalisation, which commits the audio sin of 'gain followed by attenuation', a sure recipe for overload problems. It may be possible to use passive equalisation with valves running on 250V supply rails, but it can cause problems with op amps running on standard rails. (On the other hand, valves present real difficulties for RIAA equalisation, since their output resistance changes as they age.)

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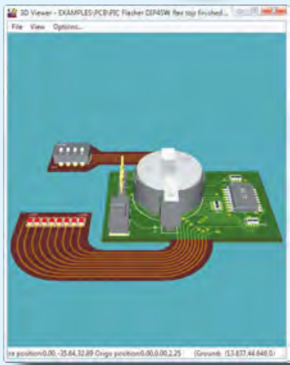

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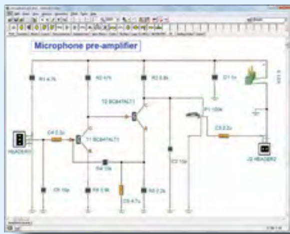

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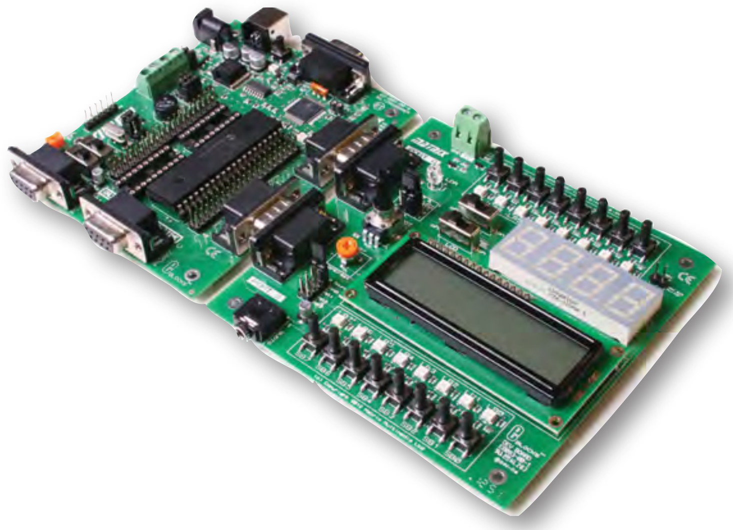

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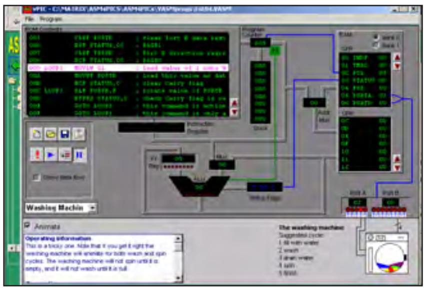

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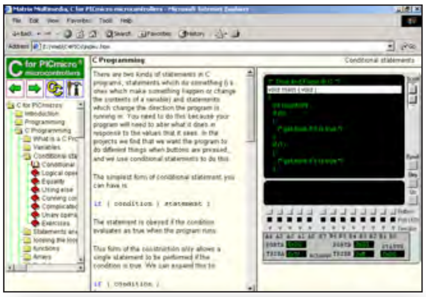

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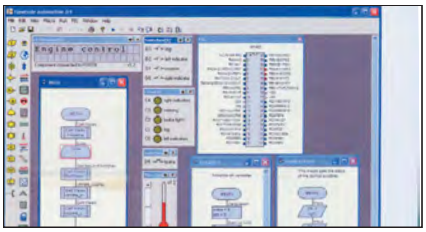

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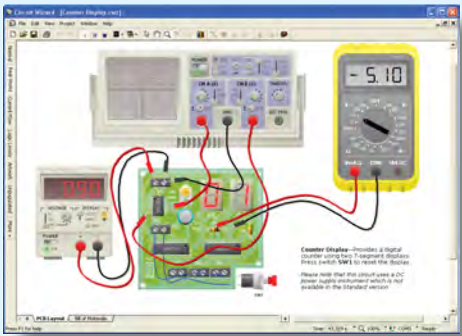

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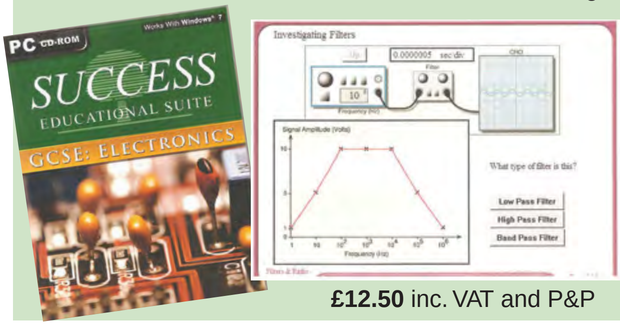

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Max's Cool Beans

By Max The Magnificent

Tri-colour LEDs – Part 2

In July's column, we introduced the concept of tri-coloured LEDs, which boast red, green, and blue (RGB) LEDs in a single package. We also noted that, if we limit ourselves to simply turning each channel (sub-LED) on or off, then we end up with 2^3 = 8 different colour combinations: black (all off), red, green, blue, yellow (red and green), magenta (red and blue), cyan (green and blue), and white (red, green, and blue).

An alternative technique is to vary the brightness of the channels, in which case we can potentially generate millions of colours. But how are we to vary the brightness? Well, we could vary the current by adjusting the resistor (unusual, but possible), or we could vary the drive voltage (perhaps by using a digital-to-analogue converter). If these techniques are performed correctly, the brightness can be controlled all the way down to dimmer than the human eye's response, but there are some downsides. For one, the colour (or wavelength) shifts over the dimming range by a very perceptible amount for most LED types. Also, the current at which the LED cuts off (no light output) varies across manufacturing lots (silicon/doping variations, temperature etc), which makes each device somewhat unpredictable at the low-output end. Furthermore, the light output can vary as a function of temperature when using these analogue dimming approaches.

The most commonly used solution is to simply turn the LEDs on and off very quickly. By varying the amount of time they are on compared to the amount of time they are off, we can effectively control their brightness without any of the problems associated with low-current response.

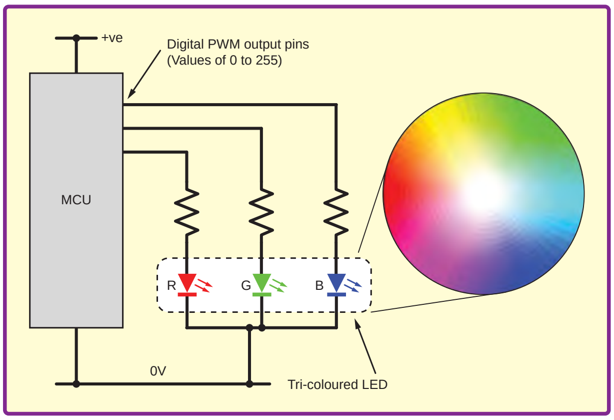

Fig.2. Controlling a tri-coloured LED using three of the MCU's PWM outputs

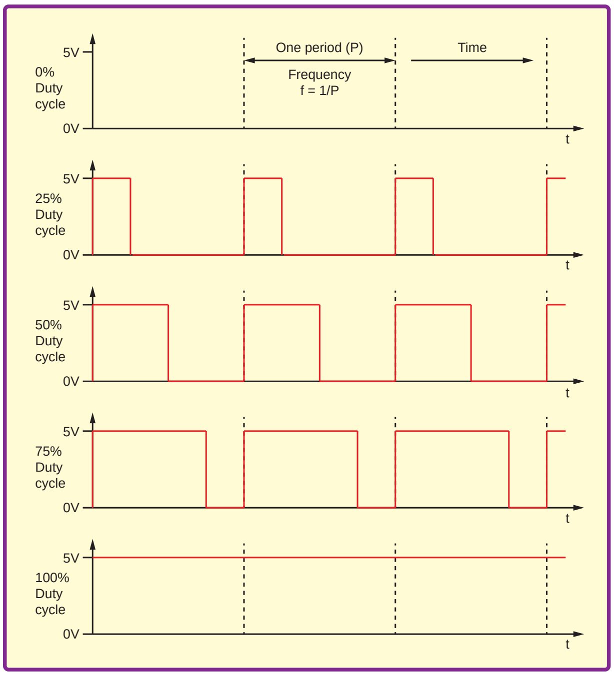

Fig.1. Pulse-width modulation (PWM)

Pulse-width modulation (PWM)

Although this is really not too complicated, it can be a tad tricky for beginners to wrap their brains around, so let's take things step-by-step. Don't worry about how we do this for the moment – let's simply assume that we decide to drive an LED with a regular square wave in the form of a signal that varies between being off and on (0V and 5V, respectively). The term 'duty cycle' refers to the percentage of one period in which a signal is active (on). Fig.1 shows, from top to bottom, duty cycles of 0%, 25%, 50%, 75%, and 100%.

This approach is known as pulse-width modulation (PWM). In the case of a 0% duty cycle, our LED would be completely off. In the case of a 100% duty cycle, our LED would be fully on. But what about a duty cycle of 50%, for example? Well, this all depends on the frequency of the signal. If the signal had a period (P) of one second, then the frequency (f) would be given by $1/P$ = 1Hz, or one cycle per second. In this case, we would see the LED flashing on and off at the same rate as one might count 'Thousand one, thousand two, thousand three,' and so forth.

If we were to switch the LED on and off fast enough, however, then the human eye wouldn't be able to perceive any flicker, and a 50% duty cycle would equate to the LED appearing to be about half as bright as when it is fully on. Similarly, a 25% duty cycle would correspond to a dim glow; a 75% duty cycle would correspond to a medium brightness; and a 100% duty cycle would equate to the LED being fully on.

It's all about switching

The great thing about electronics is that we can switch things on and off hundreds of thousands (even millions) of times a second, if we wish. And the great thing about microcontrollers (MCUs) like the Arduino is that

they contain special PWM blocks that are associated with certain pins. In the case of the Arduino Uno, for example, the pins D3, D9, D10, and D11 have a PWM frequency of 490Hz (ie, 490 cycles per second), while pins D5 and D6 have a PWM frequency of 980Hz. With regard to driving an LED, both of these frequencies are sufficiently high that changes in the duty cycle will be perceived as variations in brightness without any apparent flickering effects.

The Arduino Uno boasts six 8-bit PWMs, which means we can assign each of them 2^8 = 256 different values ranging from 00000000 to 11111111 in binary or 0 to 255 in decimal. In order to drive one of the Arduino's PWM pins in a PWM fashion, we use the `analogWrite()` function. This accepts two arguments: the number of the pin and the required PWM value. Suppose we wish to drive pin D3 in a PWM fashion, for example, we could do so as follows:

```
analogWrite(3,0);   //   0% duty cycle
analogWrite(3,64);  //  25% duty cycle
analogWrite(3,127); //  50% duty cycle
analogWrite(3,191); //  75% duty cycle
analogWrite(3,255); // 100% duty cycle
```

Tri-colour control

Now let's return to our tri-coloured LEDs. If we use three MCU PWM outputs to control the device – one output for each RGB sub-channel – then we can theoretically achieve 2^8 * 2^8 * 2^8 = 16,777,216 different colours, as illustrated in Fig.2.

Having more than 16 million colours at our fingertips affords us the ability to achieve some rather tasty effects. For example, consider the 'rainbow' effect shown at the end of this video of my *Bodacious Acoustic Diagnostic Astoundingly Superior Spectromatic* (*BADASS*) display (**http://bit.ly/1EAzRbD**). And here's another video that shows the display responding to music (**http://bit.ly/1FQm0TW**).

The problem with the scheme illustrated in Fig.2 is that each tri-coloured LED requires three MCU pins. If we were to use this technique to drive my BADASS display, which boasts an array of 16 x 16 = 256 pixels / elements, we would need an MCU with 3 * 256 = 768 PWM-enabled pins. In fact, I can drive my entire display using only a single Arduino Uno pin, if I so desire. How is this possible? All will be revealed in my next *Cool Beans* column. Until next time, have a good one!

One last thing...

Before I forget, fancy a different flavour of 'beans'? Next month, check out my new *Hot Beans* blog!

Any comments? – please feel free to email me at: **max@CliveMaxfield.com**.

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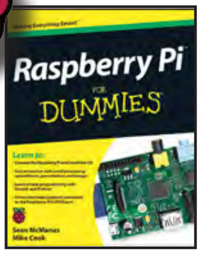

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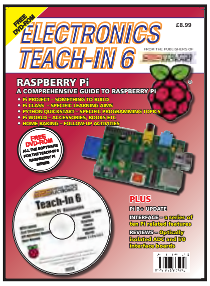

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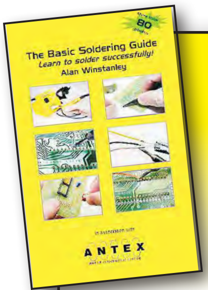

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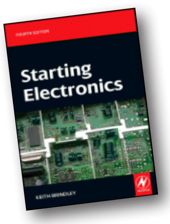

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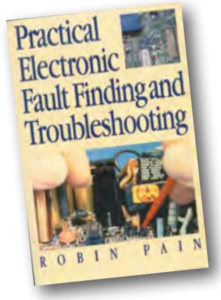

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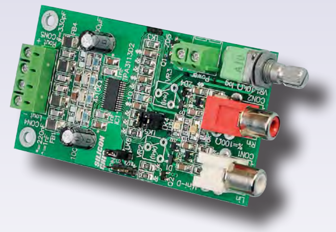

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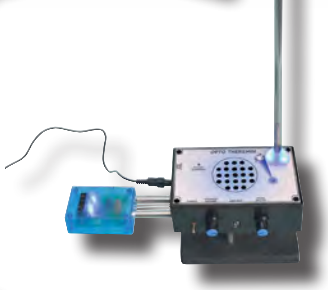
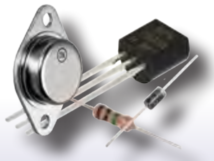


Wideband, active differential oscilloscope probe

Using your oscilloscope to examine and measure high-speed and high-frequency circuits can be tricky if you only use the usual passive test probes supplied. Here's a design for a high performance, active differential probe that costs much less than commercially available active probes. It has very little circuit loading and a useable bandwidth of more than 80MHz.

Teach-In 2015 – Part 8

In September's *Teach-In 2015*, we'll examine high-power amplifiers, and the Darlington and Sziklai pair configurations that are commonly found in them. Negative feedback will be introduced and we'll explain how it provides a useful and very effective way of making an amplifier stable and predictable.



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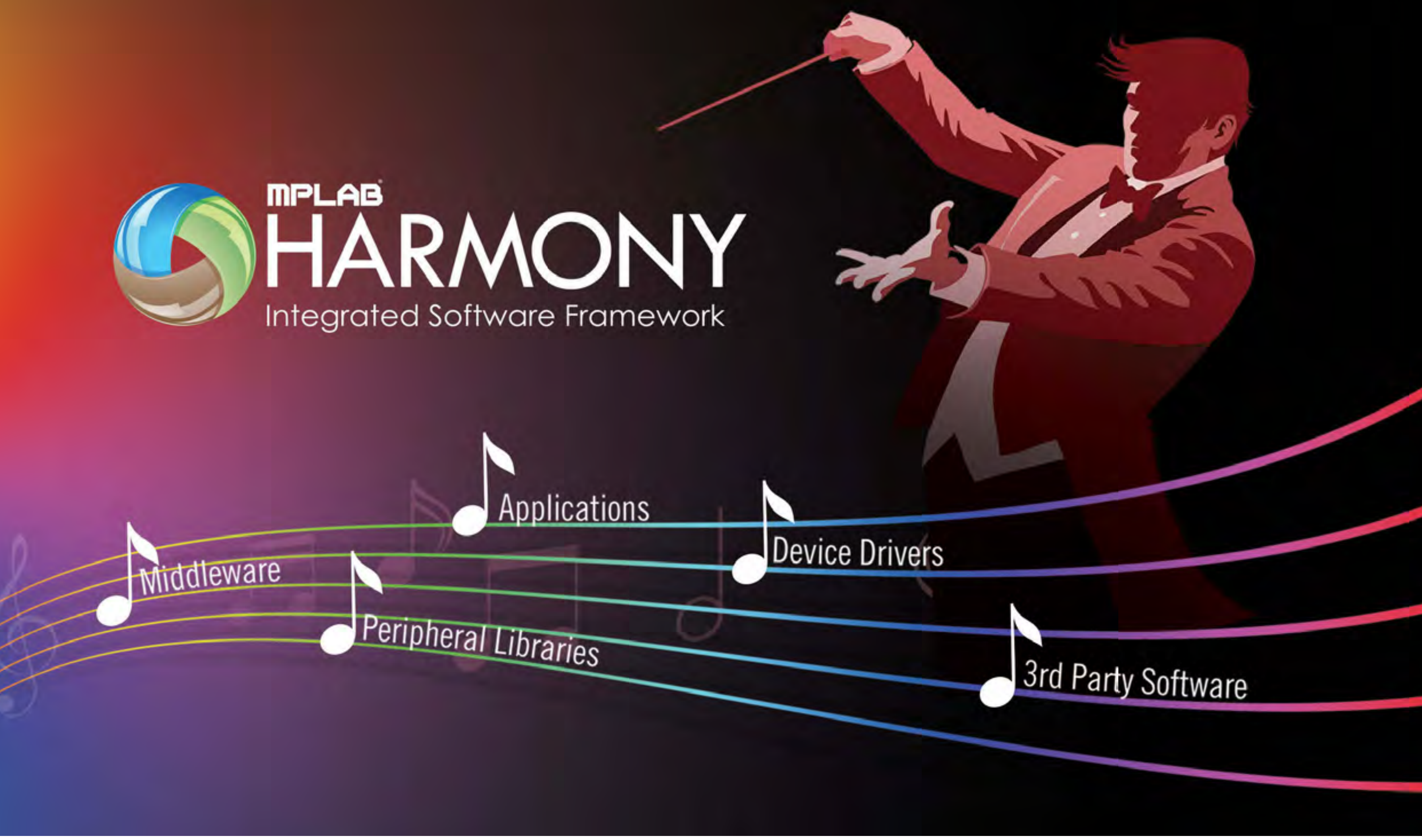
Faster PIC32 Development with Fewer Resources



Code Interoperability

- Modular architecture allows drivers and libraries to work together with minimal effort

Faster Time to Market

- Integrated single platform enables shorter development time

Improved Compatibility

- Scalable across PIC32 Microchip parts to custom fit new project requirements

Quicker Support

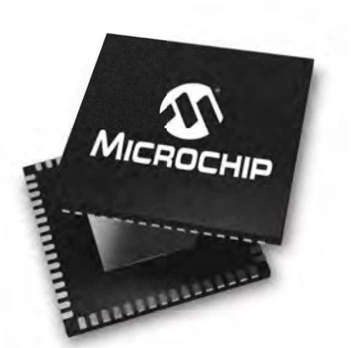
- One stop support for all of your design needs, including third party solutions

Easy Third Party Software

- Integrates third party solutions into the software framework seamlessly

GUI Project Confi guration

- Fast, accurate project creation and configuration, including third parties



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